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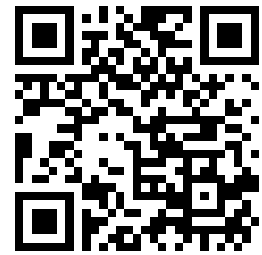

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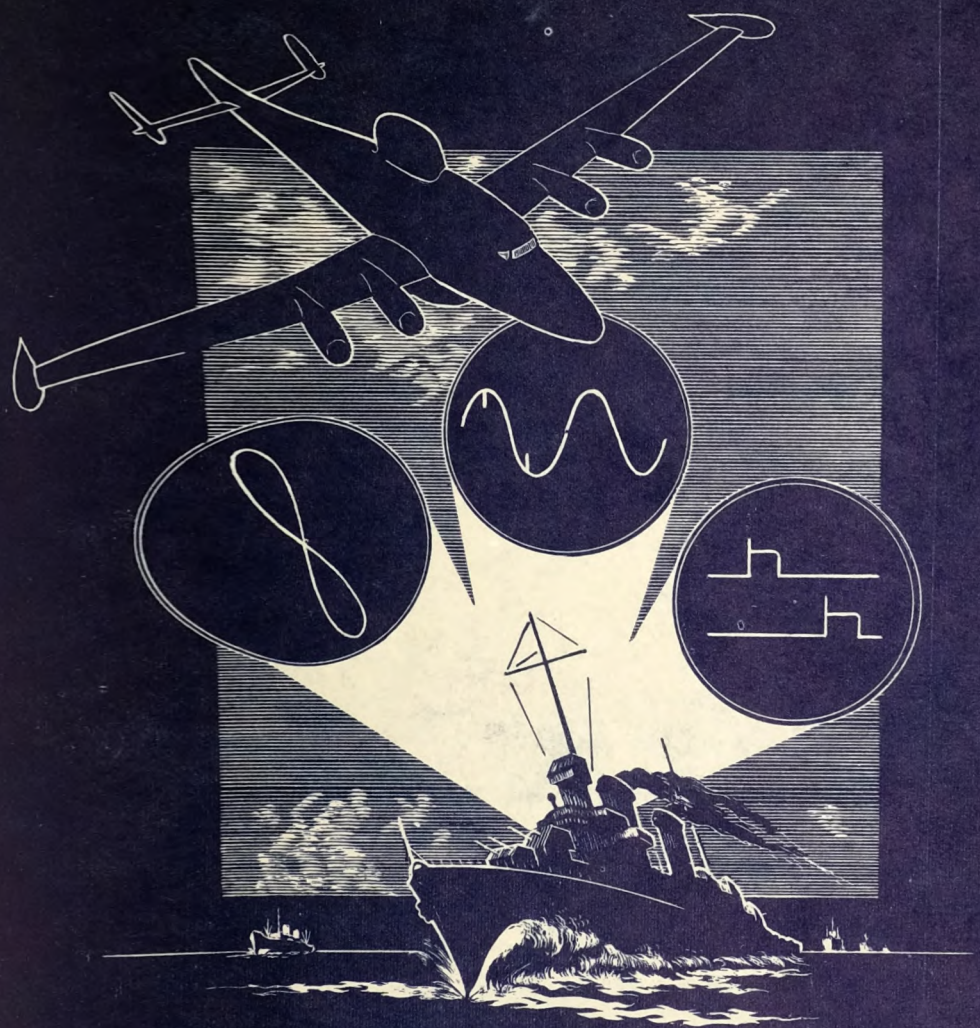

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ELECTRONICS TECHNICIAN 2, VOL. 1

BUREAU OF NAVAL PERSONNEL

NAVY TRAINING COURSE NAVPERS 10190-B

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PREFACE

Electronics Technician 2, Volume 1, is one of a two-volume series written as an aid to enlisted men who are preparing for advancement to ET 2. This book is one of a series of Navy Training Courses designed to give enlisted men the background material necessary to perform the duties of their rate.

This training course contains information about test equipment, transmitters, receivers, teletype, single sideband, Loran, Tacan, and radiac needed by the Electronics Technician in performing his duties within the rate of ET 2.

As one of the Navy Training Courses, this book was prepared by the U. S. Navy Training Publications Center under the supervision of the Bureau of Naval Personnel.

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THE UNITED STATES NAVY

GUARDIAN OF OUR COUNTRY

The United States Navy is responsible for maintaining control of the sea and is a ready force on watch at home and overseas, capable of strong action to preserve the peace or of instant offensive action to win in war.

It is upon the maintenance of this control that our country's glorious future depends; the United States Navy exists to make it so.

WE SERVE WITH HONOR

Tradition, valor, and victory are the Navy's heritage from the past. To these may be added dedication, discipline, and vigilance as the watchwords of the present and the future.

At home or on distant stations we serve with pride, confident in the respect of our country, our shipmates, and our families.

Our responsibilities sober us; our adversities strengthen us.

Service to God and Country is our special privilege. We serve with honor.

THE FUTURE OF THE NAVY

The Navy will always employ new weapons, new techniques, and greater power to protect and defend the United States on the sea, under the sea, and in the air.

Now and in the future, control of the sea gives the United States her greatest advantage for the maintenance of peace and for victory in war.

Mobility, surprise, dispersal, and offensive power are the keynotes of the new Navy. The roots of the Navy lie in a strong belief in the future, in continued dedication to our tasks, and in reflection on our heritage from the past.

Never have our opportunities and our responsibilities been greater.

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ACTIVE DUTY ADVANCEMENT REQUIREMENTS

REQUIREMENTS *	E1 to E2	E2 to E3	E3 to E4	E4 to E5	E5 to E6	E6 to E7	† E7 to E8	‡ E8 to E9
SERVICE	4 mos. service—or completion of recruit training.	6 mos. as E-2.	6 mos. as E-3.	12 mos. as E-4.	24 mos. as E-5.	36 mos. as E-6.	48 mos. as E-7. 8 of 11 years total service must be enlisted. Must be permanent appointment.	24 mos. as E-8. 10 of 13 years total service must be enlisted.
SCHOOL	Recruit Training.		Class A for PR3, DT3, PT3.			Class B for AGCA, MUCA, MNCA.		
PRACTICAL FACTORS	Locally prepared check-offs.	Records of Practical Factors, NavPers 760, must be completed for E-3 and all PO advancements.						
PERFORMANCE TEST			Specified ratings must complete applicable performance tests before taking examinations.					
ENLISTED PERFORMANCE EVALUATION	As used by CO when approving advancement.		...s toward performance factor credit in advancement multiple.					
EXAMINATIONS	Locally prepared tests.		Service-wide examinations required for all PO advancements.				Service-wide, selection board, and physical.	
NAVY TRAINING COURSE (INCLUDING MILITARY REQUIREMENTS)		Required for E-3 and all PO advancements unless waived because of school completion, but need not be repeated if identical course has already been completed. See NavPers 10052 (current edition).					Correspondence courses and recommended reading. See NavPers 10052 (current edition).	
AUTHORIZATION	Commanding Officer		U.S. Naval Examining Center			Bureau of Naval Personnel		
	TARS are advanced to fill vacancies and must be approved by CNARESTRA.							

* All advancements require commanding officer's recommendation.
† 2 years obligated service required.
‡ 3 years obligated service required.

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INACTIVE DUTY ADVANCEMENT REQUIREMENTS

REQUIREMENTS *		E1 to E2	E2 to E3	E3 to E4	E4 to E5	E5 to E6	E6 to E7	E8	E9
	FOR THESE DRILLS PER YEAR								
TOTAL TIME IN GRADE	48	6 mos.	6 mos.	15 mos.	18 mos.	24 mos.	36 mos.	48 mos.	24 mos.
	24	9 mos.	9 mos.	15 mos.	18 mos.	24 mos.	36 mos.	48 mos.	24 mos.
	NON-DRILLING	12 mos.	24 mos.	24 mos.	36 mos.	48 mos.	48 mos.		
DRILLS ATTENDED IN GRADE †	48	18	18	45	54	72	108	144	72
	24	16	16	27	32	42	64	85	32
TOTAL TRAINING DUTY IN GRADE †	48	14 days	14 days	14 days	14 days	28 days	42 days	56 days	28 days
	24	14 days	14 days	14 days	14 days	28 days	42 days	56 days	28 days
	NON-DRILLING	None	None	14 days	14 days	28 days	28 days		
PERFORMANCE TESTS				Specified ratings must complete applicable performance tests before taking examination.					
PRACTICAL FACTORS (INCLUDING MILITARY REQUIREMENTS)		Record of Practical Factors, NavPers 760, must be completed for all advancements.							
NAVY TRAINING COURSE (INCLUDING MILITARY REQUIREMENTS)		Completion of applicable course or courses must be entered in service record.							
EXAMINATION		Standard exams are used where available, otherwise locally prepared exams are used.						Standard EXAM, Selection Board, and Physical.	
AUTHORIZATION		District commandant or CNARESTRA					Bureau of Naval Personnel		

* Recommendation by commanding officer required for all advancements.

† Active duty periods may be substituted for drills and training duty.

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READING LIST

NAVY TRAINING COURSES

Electronics Technician, 3, NavPers 10188-A
Electronic Technician 2, Vol. 2, NavPers 10191A
Basic Electricity, NavPers 10086
Basic Electronics, NavPers 10087
Basic Hand Tool Skills, NavPers 10085
Blueprint Reading and Sketching, NavPers 10077-A

OTHER PUBLICATIONS

U.S. Navy Safety Precautions, chapter 18, OpNav 34P1

USAFI TEXTS

United States Armed Forces Institute (USAFI) courses for additional reading and study are available through your information and Education Officer.* The following courses are recommended:

B150 Review Arithmetic
B151 General Mathematics I
B152 General Mathematics II
C164 Beginning Algebra I
C165 Beginning Algebra II
C166 Advanced Algebra
C290 Physics I
C291 Physics II
B781 Fundamentals of Electricity
B885 Fundamentals of Radio
B890 Radio Servicing and Repair I

* "Members of the United States Armed Forces Reserve components, when on active duty, are eligible to enroll for USAFI courses, services, and materials if the orders calling them to active duty specify a period of 120 days or more, or if they have been on active duty for a period of 120 days or more, regardless of the time specified on the active duty orders."

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CHAPTER 1

INTRODUCTION TO ET 2

The advances being experienced in the field of electronics make this one of the most desirable occupations for the young Navy man of today. The bright light of fast advancements glows for the sailor who sets for himself a high goal and pursues this goal with diligence and initiative. The technician is one of the most important single links in the chain of any successful naval operation. Without effective preventive maintenance (POMSEE) and repair (corrective maintenance) of the numerous types of equipments, many other efforts, no matter how worthwhile will be wasted. You as an ET2 must share the burden of organization, training, and repair with your seniors because your responsibilities have now grown to not only the problem of personal repair accomplishments, but also that of administration.

MILITARY

Electronics technicians perform both military and professional duties. The military duties are the same as those of other petty officers, irrespective of their professional or specialty ratings.

Underway, you must man general quarters stations, which may include (1) being a member of a repair party or (2) standing watch on equipments in CIC as both operator and technician. Also as a petty officer it is your constant duty to maintain good order and discipline at all times. This includes such details as the cleanliness and organization of your living and working spaces.

In port, you may be called upon to perform additional assignments for such military duties as shore patrol, security watches, or taking charge of a draft of men to a fleet school (being responsible for their safe delivery). Military duties are not discussed in detail in this training course, but to be a top-notch petty officer as well as technician, a thorough knowledge of your military duties is essential. A more complete discussion of military responsibilities of the ET2 can be found in *Basic Military Requirements*, NavPers 10054 and in *Military Requirements for Petty Officers 3 and 2*, NavPers 10056.

Navy ships are operated under standard administrative and battle organizations. This organization divides the ship's personnel into the (1) operations, (2) navigation, (3) gunnery, (4) engineering, (5) supply, (6) medical and (7) dental departments. Aircraft carriers and seaplane tenders have, in addition, an air department, and repair ships have a repair department.

The scope of this training course does not permit a description of the entire ship's organization, which is published in *United States Navy Regulations* and more specifically in the Ship's Organization and Regulations Manual. However, a brief description of the operations department is included with particular emphasis on the electronics division and the duties and responsibilities of the Electronics Technician.

The operations department is under the direct supervision of the operations officer. The department normally consists of six divisions. On larger vessels you can expect to be placed in the OE division. On smaller vessels you might find yourself integrated into any one of the other divisions, for example, the OI division.

The maintenance and repair of all electronic equipment, except as assigned to another department, is the responsibility of the operations officer.

The electronics material officer is responsible, under the operations officer, for the organization, administration, and operation of the electronics division and its assigned personnel and material in support of the over-all mission of the ship. He is responsible for the technical maintenance and repair of all electronic equipment, the supervision and training of all ET's, the effectiveness of the POMSEE programs,

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the maintenance and submission of all necessary records and reports, and the upkeep and cleanliness of all spaces under his cognizance.

The shipboard electronics technician is usually a member of the operations department and shall be under the direct supervision of the electronics material officer (EMO).

Many times the responsibility falls upon a petty officer who is not a chief to supervise and complete many functions cognizant to the electronics division. This distinction will fall to you if you are the senior petty officer (leading petty officer) technician.

The senior electronics technician assigned to the vessel is responsible for, and acts as, an assistant to the EMO. He is at all times charged with the direct supervision and execution of all of the many orders and repairs involved in the electronics mission. He must be alert to, and aware of, any deficiency in the electronic system and be in constant contact with all the men under his authority. His responsibility is to insure proper compliance with all of the orders of the EMO.

Due to the complexities and vast numbers of equipments in certain categories, the repair, upkeep, and preservation in these categories should be made into specialty responsibilities. A convenient grouping for assignment might be (1) Communications Electronics Technician, (2) Radar Electronics Technician, and (3) Auxiliary and Miscellaneous Electronics Technician.

The general specialty duties of ET's are to maintain and repair all electronic equipment. This equipment includes radio; radar; electronic countermeasures; and other types of communications, detection, and ranging equipment employing electronic circuits. These duties do not include aircraft electronic equipments and ordnance electronic equipments.

The Electronics Technician must also be able to use special testing equipment, such as multimeters, vacuum-tube voltmeters, tube testers, signal generators, frequency meters, and cathode-ray oscilloscopes. He must be able to disassemble equipment and replace defective parts, and also to calibrate, tune, and adjust equipment.

Specifically, the ET2 must be able to perform the duties outlined in the *Manual of Qualifications for Advancement in Rating,* NavPers 18068.

On some small ships that do not have sufficient assigned storekeepers to administer the electronic maintenance parts, this responsibility may be a requirement of the electronics division. In such cases, the EMO will assign this responsibility to an ET as a collateral duty.

MAINTENANCE

To attain the highest degree of operational and material readiness, all electronic material under the technical control of the Bureau of Ships must be maintained in a satisfactory operating condition, to be capable of rendering its designed performance. This performance must be as prescribed in the applicable performance standard (POMSEE), or technical manual if the former is not available. In furthering this policy for equipment and systems, maintenance considerations start with the basic design of an electronic piece of equipment and continue until disposed of.

MAINTENANCE is the function of retaining material in, or restoring it to, a serviceable condition. Its phases include servicing, repair, modification, modernization, overhaul, rebuild, test, reclamation, inspection, condition determination, and the providing of support items. This is accomplished under several main categories:

OPERATIONAL maintenance consists normally of inspection, cleaning, servicing, preservation, lubrication, and adjustment as required, and may also consist of minor parts replacement not requiring technical skills, or internal alignment.

TECHNICAL maintenance will normally be limited to maintenance consisting of replacement of unserviceable parts, subassemblies, or assemblies and the alignment, testing, and adjustment (internal) of equipment. This work requires skill and detailed knowledge of equipment.

PREVENTIVE maintenance is the systematic accomplishment of items deemed necessary to reduce or eliminate failure, and prolong the useful life of the equipment. These items are more specifically defined and outlined by the technical manual furnished with each equipment. This work requires skill and detailed knowledge of equipment.

CORRECTIVE (repair) maintenance is the correction of damage incurred through long use, accident, or other causes.

The POMSEE program (discussed in more detail later in this volume) is the official method by which the Bureau of Ships promulgates performance standards and standardized test

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procedures, and serves as the basis for a recommended preventive maintenance program for all shipboard electronic equipment under the technical control of the Bureau of Ships. POMSEE measurements are accomplished aboard ship by the maintenance activities, as required, even though the ship may not have the required test equipment.

PURPOSE AND CONTENT OF TRAINING COURSE

Science and technology have made possible many new and varied equipment developments in the field of electronics; this has given our Fleet accurate fire power and mobility. These new system developments are often extremely complex. Because of this complexity of electronic equipment, you must have a good working knowledge of the basic principles of electricity and electronics as well as a working knowledge of practical mathematics. Sources of information on these subjects are listed later in this chapter.

This course is written to cover the knowledge factors and, insofar as practicable, the PRACTICAL FACTORS up to and including change No. 15 in the Electronic Ratings (group III) of the *Manual of Qualifications for Advancement in Rating,* NavPers 18068.

The increasing number of essential equipments in use by our modern Navy necessitates that the material coverage for the ET2 training course be in two volumes. These two volumes will provide information so that you, as the prospective 2nd class, can intelligently pursue practical on-the-job repairs, and develop and demonstrate this skill with authority and a high degree of assurance.

Many of your advancement examination questions will be taken directly from the material contained in these volumes. Volume 1 contains a broad coverage of the communications field and its associated terminal equipments, including the theory of operation and its problems. Volume 2, in like manner, covers the radar field, including the theory of operation and its problems. Sonar has been excluded from this training course since the repair responsibility of this type equipment has been delegated to the Sonarman rate.

HOW TO USE A TRAINING COURSE

Generally, methods of study are the same for everyone, but the real art entails discovery of the methods that are best suited for each individual. If an equipment is available, the student should pursue his study, using the technical manual for the equipment, and relate the physical location and size of each component to that in the text or study guide. When possible, it is best to study about a specific equipment while working on it. On the job, the trainee should learn by doing.

When studying theory or operational material, some definite study plan must be established. Therefore, study must be thought of and become a habit. The material should be read and studied with as much concentration as possible. Therefore, select your conditions and surroundings so that you will not be disturbed or distracted.

It is important that learning be done in an orderly fashion so that acquired bits of knowledge will serve as stepping stones in the learning process. Technical matter should not be read with the idea of covering a specific number of pages or chapters without regard to the complexity of the subject matter. It is far better to read a small amount of material and digest it thoroughly than to cover a large number of pages and have only a vague idea of what has transpired.

Basic material should be read in order to get a thorough background before proceeding to more difficult material. It is easier to grasp new concepts with a solid basic background of fundamentals. DO NOT SKIP OVER THE FUNDAMENTALS.

When basic principles are not mastered they may, at a later date, block the mental processes needed to understand electronic circuit operation, or better, why it will not operate.

REFERENCE MATERIAL

The technician faces a burden of excessive numbers of electronics reference materials. The main problem now lies in selecting the proper type of reference material, which is best suited to the needs of the trainee and the particular equipment or job.

Reference material includes texts, handbooks, bulletins, installation, maintenance, and training manuals, instruction books, and technical publications. In addition to these service references, there are many useful commercial texts and books that may be used.

The Electronics Technician 2 in preparing for advancement in rate must study certain

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publications in addition to this training course. The Reading List in the front of this book is especially useful as supplementary study material. The references listed under the headings, Navy Training Courses and Other Publications, are of particular importance because questions on the examination for advancement may be based on material contained in these courses and publications as well as on material in this training course.

The references contained in the Reading List are taken from *Training Publications for Advancement in Rating,* NavPers 10052-( ), Group III. Electronics (fig. 1-1) is an annual bibliography published by the Bureau of Naval Personnel. This bibliography lists the current Navy Training Courses and other publications that have been prepared for the use of all enlisted personnel concerned with advancement in rating examinations. It is used by the Naval Examining Center in preparing military and professional examinations for advancement in rating and also by personnel preparing to take these competitive service-wide examinations.

Navy, USAFI, and commercial (Navy sponsored) courses can be obtained by application to your Information and Education officer. He can also help you to acquire publications that may not be readily available locally.

Navy Training Courses

You as a prospective ET2 will find that certain basic Navy training courses are absolute necessities. They include *Basic Electricity,* NavPers 10086; *Basic Electronics,* NavPers 10087; *Basic Hand Tool Skills,* NavPers, 10085; *Mathematics,* Vol. 1, NavPers 10069-B; *Mathematics,* Vol. 2, NavPers 10070-B; and *Blueprint Reading and Sketching,* NavPers 10077-A.

USAFI Courses

In addition to the Navy training courses contained in the Reading List, USAFI courses should be included as supplementary study material for all technicians.

USAFI also provides many correspondence courses (too numerous to list) in the field of physics, chemistry, blueprint reading, and mathematics. USAFI courses will be very helpful for those who lack sufficient training in these fields necessary to understand the more complicated and refined electronic circuit analysis and waveforms.

ELECTRONICS TECHNICIAN (ET)	*Electronics Technician 3	NavPers 10188-A	91373-2 (†)	3
	*Electronics Technician 2, Vol. 1 (†)	NavPers 10190-B	91374-1	2
	*Electronics Technician 2, Vol. 2 (†)	NavPers 10191-A	91375	2
	*Electronics Technician 1 & C	NavPers 10192	91376	1, C
	Basic Electricity	NavPers 10086	91224	3
	Basic Electronics	NavPers 10087	91227	3
	U.S. Navy Safety Precautions (Chapter 18)	OpNav 34P1		3
	BuShips Technical Manual	Chapter 60, Part 6		2
	BuShips Technical Manual	Chapters 67, 69		1

(†) When available, about July or August 1960.

Figure 1-1. —Training publications for advancement in rating (group III).

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Other Publications

The equipment manufacturer is required to provide two instruction books with each piece of electronic equipment. These books are very important because they provide the technician sufficient information for the proper operation and maintenance of the equipment.

Instruction books are generally divided into eight sections (sometimes several volumes) in the following manner: (1) general description, (2) theory of operation, (3) installation and initial adjustments, (4) operation, (5) operator's maintenance, (6) preventive maintenance, (7) corrective maintenance, and (8) parts lists. If the ET has acquired the proper background in electricity and electronics, the theory of operation and maintenance sections should not cause too much trouble. Included with some instruction books is Stock Number Identification Tables (SNITS) shown in figure 1-2, which is a very useful part of the book.

The important part that Naval handbooks play in practical every day applications is shown by their listing in ET3. Handbooks present electronics information in a practical

Figure 1-2.—Stock Number Identification, (SNIT) main section.

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form that can be used by the technician about a particular field of work, and about a particular type of equipment. Three such publications are the *Handbook of Test Methods and Practices*, NavShips 91828A; *Handbook of Naval Shore Station Electronics Criteria*, NavShips 92675; and the *Electronic Test Equipment Handbook*, NavShips 900,155.

An absolute must in any electronics publication (listing references) is the *Electronics Information Bulletin (EIB)*, NavShips 900,022A, which is published biweekly for naval electronics activities. A complete file of these bulletins must be maintained because the information contained therein is both authoritative and directive in nature. Commencing with EIB 434, these bulletins may be referenced as the authority for adoption of ideas contained therein.

Regulations governing the requirements of electronics installations and organization are specified in the Type Commanders Instructions and Directives. Generally, for example, instructions required for electronics records may be found in the 9670 series. Depending upon the type of ship these instructions and directives will define and amplify the broad electronics requirements contained in the *Bureau of Ships Technical Manual*, 250-000, chapter 67, and the *United States Navy Regulations*, 1948.

Another important instruction you will encounter onboard ship is Reporting Electronic Equipment Installations, NavShips 900,135(C), which contains complete directions for filling out the Ship Electronics Installation Record, NavShips Form 4110, as shown in figure 1-3. This instruction should be on board and the instructions therein carefully followed in preparing and submitting the NavShips Form 4110.

The importance and purpose of NavShips Form 4110 cannot be too highly stressed. It is a report of electronic equipment installations in the Fleet used by management and command personnel in planning, which includes installation, procurement, distribution, and budgetary matters involving Bureau of Ships shipboard electronic equipment. NavShips 4110's are specifically used:

1. By the Bureau as a basis for determining the equipments to be installed during overhauls and conversions, and the equipment to be procured in the future.

2. By Naval Shipyards in overhaul planning.

3. By the electronic supply office to determine repair parts allowance.

4. To prepare the basic Ship Electronic Allowance Lists, NavShips 900,115.

5. To furnish vital information for budgetary planning.

6. To prepare the Electronic Equipment Distribution Schedule, NavShips 4311.

7. To prepare classified publications for Fleet and Type Commanders as an aid in ascertaining a vessel's capabilities and shortages. These publications are classified: Compilations of Electronic Installations in the Active Atlantic Fleet, NavShips 92779; Active Pacific Fleet, NavShips 92780; Reserve Atlantic Fleet, NavShips 92900; and Reserve Pacific Fleet, NavShips 92901.

The following publications provide much general and specific material helpful to all personnel engaged in the installation and maintenance of electronics equipment.

The first and by far the most important of these manuals is the *Bureau of Ships Technical Manual*, NavShips 250,000. This manual is published for the guidance of all persons in the Naval Establishment in accordance with Article 1201, *U. S. Navy Regulations*, 1948. Instructions contained therein are binding on all persons in the Naval Establishment.

A recent reorganization of the Electronic Maintenance Book (EMB), (EIMB, change 10), is segregated into two parts; one called the Electronics Installation and Maintenance Book (EIMB), NavShips 900,000, and the other called Service Notes, NavShips 900,000-1.

The EIMB section is published for the guidance of all personnel in the naval establishment in accordance with articles 67-2 and 67-5 of the *Bureau of Ships Technical Manual*, NavShips 250,000 and the Service Notes section contains the Service Notes formerly included in the EMB.

At the present time the EIMB is in the process of being reorganized.

This reorganization of the EIMB will reduce the time required by technical personnel to properly handle electronics work and material under the cognizance of the Bureau of Ships.

A highly informative and very useful monthly publication that contains an electronics section on the latest developments and systems in the Navy is the *Bureau of Ships Journal*, NavShips 250,200. Contents of this publication is for information only, and should not be regarded as altering or superseding official regulations, orders, or directives.

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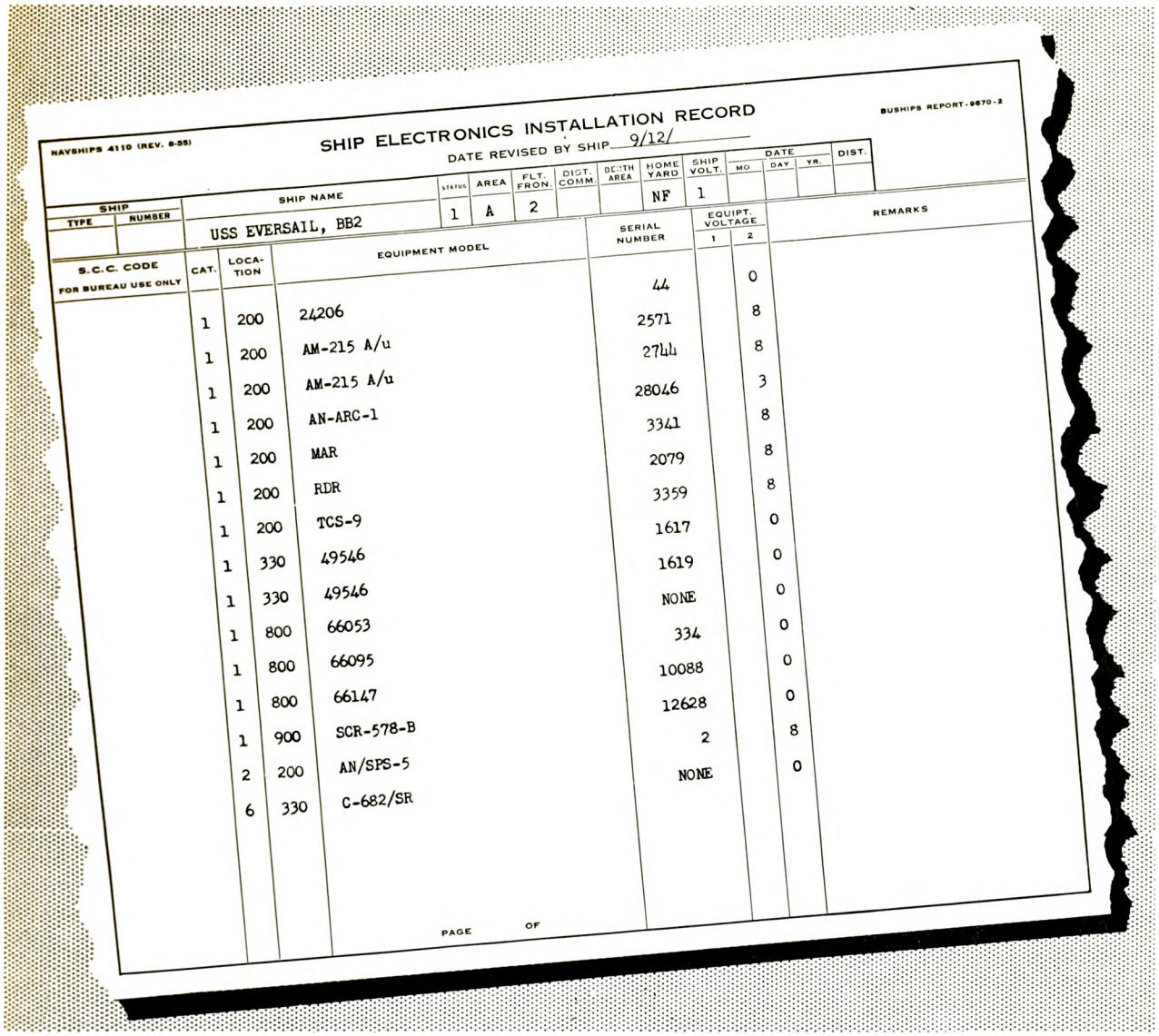

NAVSHIPS 4110 (REV. 8-55) SHIP ELECTRONICS INSTALLATION RECORD BUSHIPS REPORT-9670-2

DATE REVISED BY SHIP 9/12/

SHIP TYPE	SHIP NUMBER	SHIP NAME	STATUS	AREA	FLT. FRON.	DIST. COMM.	BERTH AREA	HOME YARD	SHIP VOLT.	DATE MO	DAY	YR.	DIST.
		USS EVERSAIL, BB2	1	A	2			NF	1				

S.C.C. CODE FOR BUREAU USE ONLY	CAT.	LOCATION	EQUIPMENT MODEL	SERIAL NUMBER	EQUIPT. VOLTAGE 1	EQUIPT. VOLTAGE 2	REMARKS
	1	200	24206	44		0	
	1	200	AM-215 A/u	2571		8	
	1	200	AM-215 A/u	2744		8	
	1	200	AN-ARC-1	28046		3	
	1	200	MAR	3341		8	
	1	200	RDR	2079		8	
	1	200	TCS-9	3359		8	
	1	330	49546	1617		0	
	1	330	49546	1619		0	
	1	800	66053	NONE		0	
	1	800	66095	334		0	
	1	800	66147	10088		0	
	1	900	SCR-578-B	12628		0	
	2	200	AN/SPS-5	2		8	
	6	330	C-682/SR	NONE		0	

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Figure 1-3.—Ship Electronics Installation Record (NavShips 4110).

There are yet many publications (too numerous to list here), which the technician may use successfully in electronic repair work. A careful study of the publications section in the *Bureau of Ships Technical Manual,* NavShips 250,000, chapter 67 will provide you with the titles of books or manuals to fit any desired subject or requirement. For example, you may need information on electronic equipment specifications and safety precautions. The manual lists (under specs) *General Specifications for Ships of the U.S. Navy* and under safety precautions, *U.S. Navy Safety Precautions,* OpNav 34P1, *Artificial Respiration,* NavMed P5002A and NavMed P5003, and *Electric Shock, Its Cause and Prevention,* NavShips 250,660-42.

The *Armed Services Index of R.F. Transmission Lines and Fittings,* NavShips 900,102(B), and the *Dictionary of Standard Terminal Designators for Electronic Equipment,* NavShips 900,186, will provide you with information about terminal designators and transmission lines.

REQUIRED KNOWLEDGE AND SKILLS

Electronics Technicians perform both military and professional duties. *The Manual of*

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Qualifications for Advancement in Rating (revised), NavPers 18068, commonly referred to as the Quals Manual, contains the military requirements and the professional qualifications for all the ratings in the Navy. The Quals Manual is periodically revised to reflect organizational and procedural changes in the Navy that affect the rating structure, and to incorporate additional skills and techniques required by the development and installation of new equipment.

A reprint of the professional qualifications for advancement of the Electronics Technician rating is presented in appendix II of this training course. These qualifications are current through change No. 14 dated 19 November 1959. Personnel preparing for any examination subsequent to the date of this change should refer to the latest revision of the Quals Manual.

IT IS MANDATORY THAT THE ET CONSTANTLY CHECK NAVPERS 18068 TO BE SURE THAT THE LATEST PERIODIC QUAL CHANGE HAS BEEN INCORPORATED.

Purpose and General Description

The enlisted rating structure contained in the *Manual of Qualifications for Advancement in Rating (revised)*, NavPers 18068, is the primary means for the classification of Navy enlisted personnel. It is a systematic alignment of occupational groups (rates and ratings) where the rating structure is subject to a continual review to ensure the most effective manpower utilization and career patterns.

A Navy Enlisted Classification Code is assigned to each technician according to his field specialty. For the Navy enlisted classification codes included within this rating, see *Manual of Navy Enlisted Classifications*, NavPers 15105-A (revised), codes ET-1500 to ET-1599.

Changes

The enlisted rating structure is the primary method for the classification of Navy enlisted personnel, and therefore is subject to a continual review to insure its most effective use. In order to meet this constant demand over the years, modifications and changes had to be effected in the form of periodic manual changes. These periodic changes do NOT necessarily affect all groups or ratings in a group.

RECORD OF PRACTICAL FACTORS

It is foolish to expect a technician to become proficient from reading and studying equipment instruction books and training manuals only. Therefore, the technician desiring to qualify for the next higher rate must obtain a certain amount of practical experience. As the candidate gains this experience and demonstrates a proficiency in each practical factor, a record must be kept.

A form, Record of Practical Factors, NavPers 760 (ET) shown in figure 1-4, is provided by the Bureau of Naval Personnel on which the supervising officer can date and initial each practical factor (professional and military) as it is completed by the candidate for the next higher rate. An Electronics Technician qualifying for the next higher rate must satisfactorily complete all of the military and professional practical factors before he can be recommended to take the service-wide examination for advancement in rating.

IT IS VERY IMPORTANT THAT THE RECORD OF PRACTICAL FACTORS CORRESPONDING TO THE LATEST QUALS MANUAL BE USED.

The Record of Practical Factors provides a standard checkoff list of the practical factors (both military and professional) required to be demonstrated in each rate as a prerequisite for advancement. When a technician demonstrates his proficiency in each practical factor listed on the form, the division officer or supervising petty officer initials and enters the date of completion in the appropriate column provided on the form.

Each division maintains a Record of Practical Factors for each enlisted man in pay grades E-2 through E-6 (apprentice through first class). When an enlisted man is transferred, the signed copy of this form is inserted in the correspondence side of the Enlisted Service Record, which is forwarded to the man's new duty station. In this way, the man's record is kept up to date and used on a continuing basis as he progresses in his rating.

IT IS VERY IMPORTANT THAT ALL DIVISION OFFICERS AND SUPERVISING PETTY OFFICERS USE AND KEEP THE RECORD OF PRACTICAL FACTORS UP TO DATE, BECAUSE EACH MAN MUST BE QUALIFIED AND SHOW A SERVICE RECORD PROOF OF THIS QUALIFICATION BEFORE HE IS ELIGIBLE TO COMPETE FOR THE NEXT HIGHER RATE.

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RECORD OF PRACTICAL FACTORS
NAVPERS 760 (ET) (Rev. 11-58)

INSTRUCTIONS

ACTIVE DUTY PERSONNEL

1. *As proficiency in each practical factor listed here is demonstrated, an entry is to be made in the DATE and INITIALS columns by the supervising officer.*
2. *Waivers may be granted for specific qualifications in accordance with current BuPers Instructions.*
3. *If a man demonstrates proficiency in skills considered to be within the scope of his rating but not listed in this list of minimum qualifications, an appropriate entry should be made on this form in the spaces provided unless the information is entered elsewhere in the man's service record.*
4. *A copy of this form is to be held by the division officer or by the appropriate supervising officer of each man in pay grade E–3 through E–6.*
5. *Upon transfer of an enlisted man, the supervising officer's copy of the form is to be signed, inserted in the correspondence side of the enlisted service record, and forwarded.*
6. *As changes in the rating structure and major changes in the Manual of Qualifications for Advancement in Rating, NAVPERS 18068, occur, new forms will be made available. Minor changes in NAVPERS 18068 should be recorded in existing forms in the spaces provided.*
7. *One copy of the printed form should be made available to each man for his personal record and guidance.*

USNR (INACTIVE DUTY) PERSONNEL

1. *Reserve Unit will enter following letters as appropriate in USNR column opposite appropriate qualification prior to trainee's commencing active duty for training:*

   Letter "A" *to indicate that classroom instruction has been given, but practical checkout is desired.*

   Letter "B" *to indicate that classroom instruction has not been given, but instruction and practical checkout are desired.*

   Qualifications unmarked should be ignored during active duty training periods.
2. *In "REMARKS" section on reverse of form, the ship, station, or unit providing training will enter explanatory notes concerning types of equipment used, extent, and frequency of training received, and similar data whenever appropriate.*
3. *Ship or station supervising officer must sign form for period of active duty for training.*

RATING	ABBREVIATION	NAME	SERVICE NUMBER
ELECTRONICS TECHNICIAN *(Through Change 11 to NAVPERS 18068)*	**ET**		

DATES OF COMPLETION OF PRACTICAL FACTORS CHECKOUT FOR RATE LEVEL:

3 _____ 2 _____ 1 _____ C _____

USNR (INACTIVE DUTY) PERSONNEL

RESERVE UNIT NO. _____ LOCATION _____

PRACTICAL FACTORS	RATES	USNR	COMPLETED DATE	COMPLETED INITIALS
MILITARY REQUIREMENTS FOR ENLISTED PERSONNEL				
100 PRACTICAL FACTORS				
101 OPERATIONAL				
20. Demonstrate how to use: (Men only.) a. Portable CO_2 extinguishers. b. Hoses, nozzles, and adapters. c. Safety lines and signals.	3			
102 MAINTENANCE AND/OR REPAIR				
5. Field strip, clean, and assemble the service pistol. (Men only.)	3			
103 ADMINISTRATIVE AND/OR CLERICAL				
1. Prepare an indoctrination schedule for a new recruit reporting for duty.	2			
2. Select and organize appropriate subject matter and instruct by the Showing (demonstration) method.	2			
3. Prepare a standard Navy letter.	2			
4. Prepare a detailed work assignment for men in your division.	1			
5. Teach a group, observing the following steps in developing the lesson: a. Setting the objectives. b. Presenting the subject matter. c. Providing trainee application through practical work and drill. d. Summarizing key points. e. Testing trainee achievement.	1			
6. Prepare and administer a written test which includes the following types of questions: a. Essay. b. Multiple choice. c. True-false. d. Completion.	1			
7. Use the following training aids and devices: a. Training films, slides, and transparencies. b. Charts, posters. c. Models and mockups. d. Demonstrators and trainers.	1			
8. Prepare a preventive maintenance schedule for an item of machinery or equipment in your division, considering possible changes to ship employment schedule.	C			
9. Demonstrate ability to conduct instruction by each of the following methods, using subject matter appropriate to each method: a. Telling (lecture). b. Questions and discussion. c. Drill and practical work. d. Written study materials.	C			
10. Plan and supervise on-the-job training program based on: a. Level of trainees' knowledge and skill. b. Degree of skill to be developed. c. Available equipment and training aids. d. Qualifications for Advancement in Rating, NavPers 18068 (Revised).	C			
104 MILITARY CONDUCT				
7. Command a squad in close-order drill.	3			
8. Fire service pistol, observing safety precautions. (Men only.)	3			
9. Relieve a watch armed with pistol. (Men only.)	3			
PROFESSIONAL QUALIFICATIONS FOR ET				
100 PRACTICAL FACTORS				
101 OPERATIONAL				
1. Manipulate external controls; read and interpret dials, meter indications, and cathode ray tube presentations on electronic equipment.*	3			
2. Utilize distribution patching system for radio transmitters, receivers, and antennas for all local and remote operating positions.	3			
102 MAINTENANCE AND/OR REPAIR				
1. Demonstrate under simulated conditions the rescue of a person in contact with an energized electrical circuit, resuscitation of a person unconscious from electrical shock, and treatment for electrical and acid burns.	3			
2. Demonstrate, while servicing equipment, safety precautions such as tagging switches, removing fuses, grounding test equipment, and using shorting bars and rubber mats.	3			

* NOTE.—Denotes items which, when applied to emergency service ratings, are limited to the equipment, tools, and test equipment included within the scope of those ratings.

Figure 1-4.—Record of practical factors.

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As changes are made periodically to the *Manual of Qualifications for Advancement in Rating (revised)*, NavPers 18068, the Record of Practical Factors (rev. 11-58), NavPers 760, (ET), may be revised. Space is allowed on the form for entering additional factors as they are published in changes to the Quals Manual. Major changes in the Quals Manual will result in a revision of the NavPers 760 (ET) form. Minor changes in NavPers 18068 should be recorded in the spaces provided on the existing forms.

ANNUAL BIBLIOGRAPHY

A very important bibliography published by the Bureau of Naval Personnel is the *Training Publications for Advancement in Rating*, NavPers 10052-( ). This bibliography lists the Navy Training Courses and other publications prescribed for the use of all personnel concerned with advancement in rating, training, and examinations.

As new Navy Training Courses and correspondence courses become available, announcements concerning them are made in the *Naval Training Bulletin*, *All Hands*, and the *Naval Reservist*.

TRAINING FACILITIES FOR ADVANCEMENT

The Navy places great emphasis on electronics training, both in the formal instruction of the service schools and in on-the-job training on ships and shore stations.

There is no real separation between training and doing in the Navy. There are jobs to do, and training consists of teaching and learning how to do the jobs. To make sure that the training is done quickly, surely, and economically, the Navy has established responsibilities, schools, study courses, and clear-cut standards and requirements to follow.

To meet the training need for self-study, a program of training and correspondence courses, discussed earlier in this chapter, is provided by the Navy. Further, to carry out the Navy training program, service schools have been established, using teaching practices, which conform to those that have proven successful over the years in civilian vocational education. Also the Navy relies heavily on the top rated personnel to teach technical skills in practical on-the-job situations.

General information on training methods and techniques is included in the *Manual for Navy Instructors*, NavPers 16103-B; and in the *Military Requirements for Petty Officers 3 and 2*, NavPers 10056.

SCHOOLS

The enlisted technical schools provide training for men as individuals. The Naval schools train and prepare selected officers and enlisted personnel, including the Naval Reserve, for usefulness afloat in designated specialties and to supplement training afloat by providing advanced and/or specialized training when such training can be more advantageously given ashore. These schools are divided into three main groups: service schools, Fleet schools, and manufacturers schools.

Enlisted service schools and courses are designed and maintained to assist the forces afloat by giving instruction and training, which, because of the time allowed and facilities available, can be best given ashore.

ON-THE-JOB TRAINING

The best ships and equipment are of little value without trained personnel. The operational readiness of every unit in the Navy depends, in a large measure, on the knowledge and skills of every crew member. The Navy, therefore, places great emphasis on training, and recognizes the need for supplementing the formalized instruction of the service schools with on-the-job shipboard or shore-based training and instruction.

Instructing and training subordinates is one of the principal responsibilities of a leading ET. The extent to which he fulfills this responsibility will determine his ability as a supervisor and leading petty officer. He must impart to his subordinates much of the knowledge and skills that he acquired through the instruction and training of his supervisors.

Planned instruction means teaching a subject in an orderly step-by-step fashion. Because the instructor is dealing with individuals who possess varying degrees of skill and training, the problem is more complicated than if all the men were beginning at the same level.

When the ET2 reports for duty at a new ship or station, he should expect to find an electronics training program set up and working. He should make sure that he is prepared to carry out the training plan established by the

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EMO and to add improvements by way of suggestions. Many times, of course, emergency drills, heavy work loads, and/or other distractions will often disrupt the schedule. Whenever possible this lost time should be made up.

Much of the on-the-job training is directly responsible for the practical factors, and the instruction obtained leads to the completion of the practical and examination factors listed in the Quals Manual. Since the majority of men are in the general-duty status, this training will provide the major source of rate training.

The most helpful text to use in setting up this program is the NWP50. Also a more detailed discussion of this type training is contained in *Electronics Technicial 1 and C,* NavPers 10192, including a list of sound picture films as found in *United States Navy Film Catalog,* NavPers 10000, and its latest supplements.

SUPERVISION AND LEADERSHIP

As you advance in rating you will assume increased responsibility for supervising, instructing, and training lower rated men and strikers. The techniques for handling men require practice and experience just as your proficiency in an occupational skill requires actual on-the-job experience.

QUIZ

1. Can an ET, other than a chief, be held responsible directly for the organization and supervision of the electronics division?
2. Name the three suggested electronics groups for the repair, upkeep, and preservation of all electronic equipment.
3. What type of maintenance consists of replacement of unserviceable parts and testing and adjustment (internal) of equipment?
4. What type of maintenance consists of a systematic work program to eliminate or reduce failures and prolong useful equipment life?
5. How should the trainee develop a thorough background in electronic fundamentals before proceeding to more advanced materials?
6. Which biweekly electronics publication can be used as an authority for performing the suggested ideas contained therein?
7. What is the name of the form that is provided by the Bureau on which the supervisor can check and initial each practical factor as the candidate completes it?
8. Which form must be kept up to date and filed at all times in each man's service record to prove his qualifications when competing for his next higher rate?
9. What are the three types of schools provided by the Navy for training men in all rates and ratings?
10. Training that is directly responsible for the practical factors and instruction leading to the completion of the practical and examination factors is known as what?
11. What type of training is considered to be the major source of rate training?

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Chapter 2

ELECTRONIC AIDS TO NAVIGATION: RDF (PART I)

INTRODUCTION

Essentially, the radio direction finder (RDF) is a sensitive receiver to which a directional antenna is connected. Radio direction finding is used primarily in air navigation. It is also used in the location of personnel afloat in liferafts or lifeboats, provided these craft are equipped with a radio transmitter, and that the RDF equipment is capable of receiving the transmitted frequencies.

Radio Direction Finder, AN/URD-4 (fig. 2-1) has been chosen for presentation in this chapter because its circuits are representative of all direction finding equipments, and because it is presently in wide usage throughout the Navy. The frequency range of the Direction Finder AN/URD-4 (225 to 399.9 mc) does not permit its use with naval distress frequencies.

SYSTEM BLOCK DIAGRAM

Direction Finder, AN/URD-4, produces a visual indication of the bearing of a received signal. Fundamentally, this is accomplished by means of an indicator unit, which presents an oscilloscopic display of the output of a receiver. The equipment is designed for either shipboard or shore installation. A functional block diagram of the complete direction finding system is illustrated in figure 2-2.

Instantaneous visual direction finding information from r-f signals in the frequency range from 225.0 to 399.9 mc is provided by the AN/URD-4 equipment. The three major units of the direction finder are the antenna, the receiver group, and the azimuth indicator.

The direction finder set is provided with a frequency control system, which divides the frequency tuning range into high and low frequency bands. The low band controls frequencies from 225.0 to 299.9 mc. The high band controls frequencies from 300.0 to 399.9 mc.

The antenna of the AN/URD-4 direction finding system is a two-dipole vertically polarized, Adcock array, which is rotated about the vertical axis by a motor at a constant speed. The antenna is most sensitive to signals in the vertical plane and least sensitive to signals in the horizontal plane. Accurate direction finding information is provided, however, from signals polarized as much as 45° from vertical.

Each time a dipole passes in a plane with a given signal, maximum voltage is generated in the antenna. Because there are two dipoles in the antenna array, the separate dipoles pass through the maximum signal path at different times to introduce a modulated component on the received r-f signal. The modulated r-f voltage is fed to the receiving portion of the equipment.

Direction finding information is presented in the form of a propeller-shaped pattern on the face of the cathode-ray tube of the indicator (fig. 2-3, A). The pattern can be made to shift position to form a distorted figure 8 "sense bend" (fig. 2-3, B), the apex of which indicates the relative bearing of the received signal. The method of obtaining the propeller-shaped pattern and sense bend will be treated later. Directional information is displayed with a bearing accuracy of plus or minus five degrees.

Provision is made on the indicator cabinet (fig. 2-1) for headphones, which are used for aural reception simultaneously with D/F (direction finding) operation. Aural reception is possible on a-m or f-m signals only. Direction finding information is displayed on the cathode-ray tube for any received signal whether modulated or unmodulated.

The rotating portion of the antenna is coupled to the stationary portion (fig. 2-2) through a rotating capacitance joint. The energy transfer from the antenna to the receiver is accomplished through this capacitance connection.

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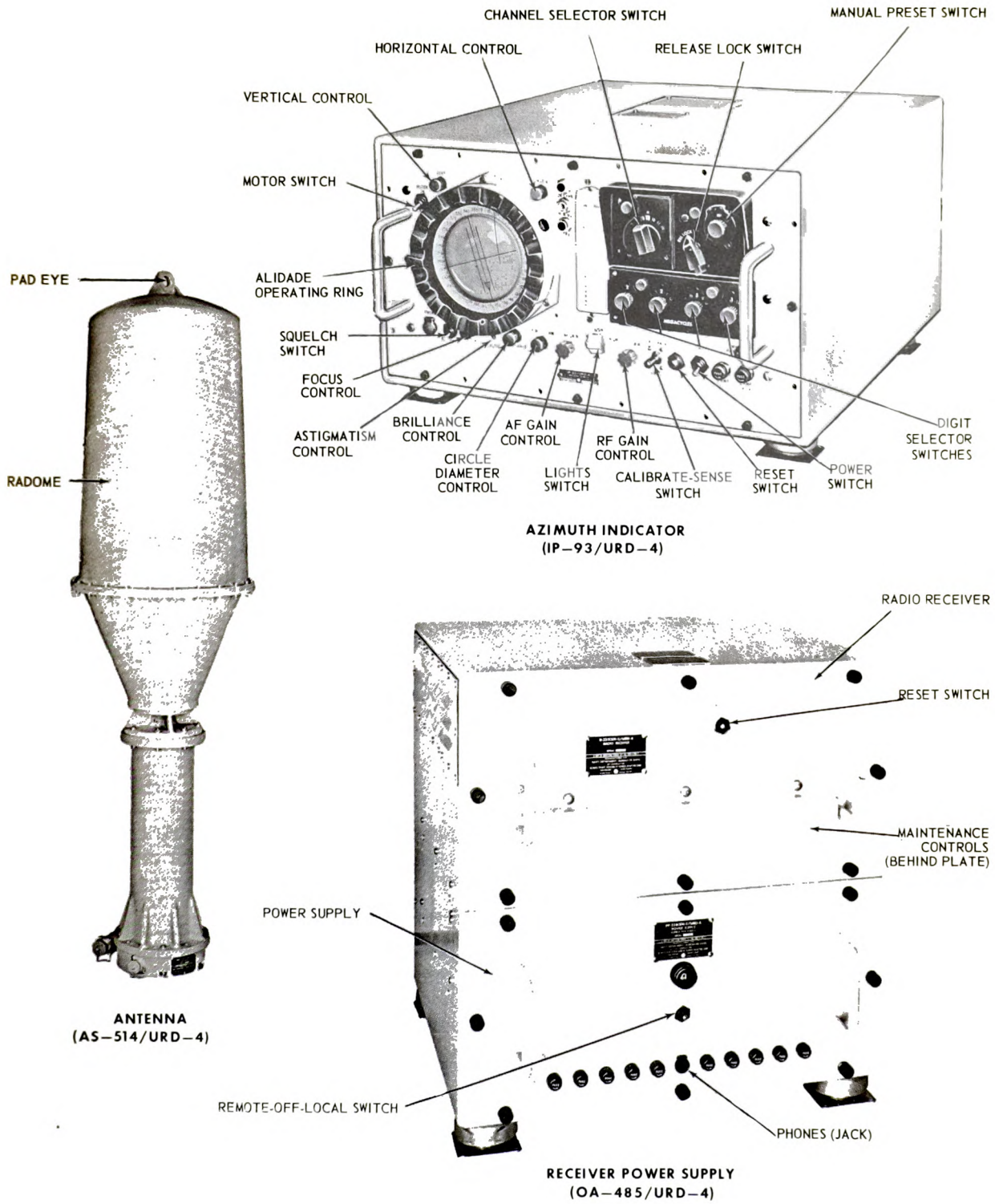

Figure 2-1. –Direction Finder Set, AN/URD-4, major components.

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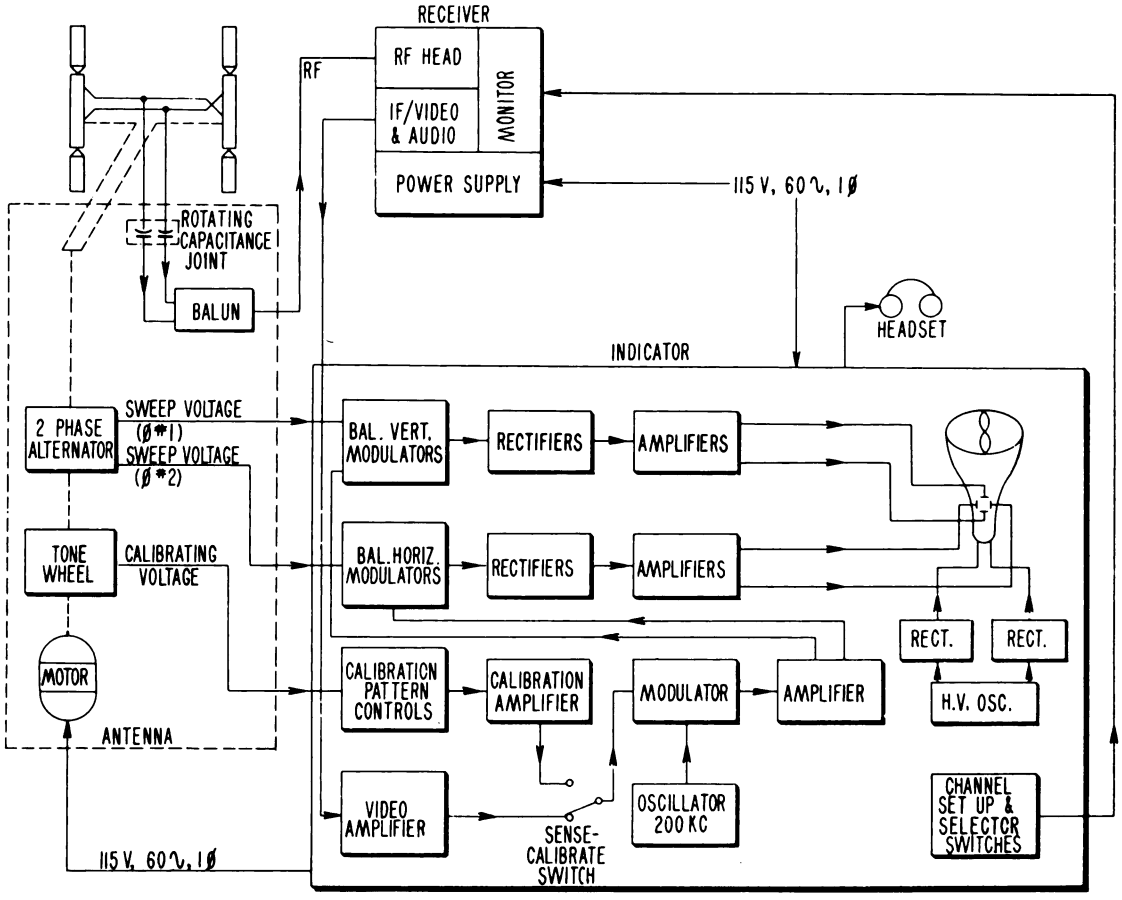

Figure 2-2. - rection finder, system block diagram.

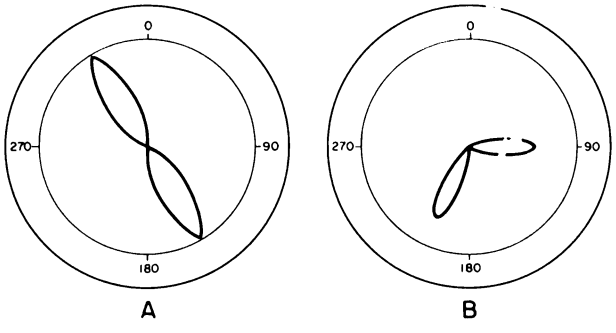

Figure 2-3. –Indicator patterns.

A two-phase alternator generates horizontal and vertical sweep voltages for application to the horizontal and vertical deflection plates of the cathode-ray tube. The phase of the two voltages are correct to form a circular pattern on the indicator. This pattern is present when no signal is being received.

Calibration voltage for calibration of the indicator unit is generated by a tone-wheel, which is connected to an alternator. The antenna array, the two-phase alternator, and the tone-wheel are all driven by the antenna drive motor.

R-f antenna voltage is fed to the receiver (fig. 2-4) through a balance-to-unbalance device referred to as a "balun." The balanced output of the antenna is unbalanced by the balun to allow r-f voltage to be transferred to the receiver by means of a coaxial cable. The r-f head of the receiver is essentially a tuned amplifier unit. The receiving frequency passed by the r-f head is determined by the monitor circuit.

The monitor controls the frequency of a variable frequency oscillator (vfo). The output of the vfo is multiplied and used as the heterodyning frequency (in the first mixer) for the signal input to the r-f head.

The monitor consists of several stages of heterodyning. It provides its controlling action

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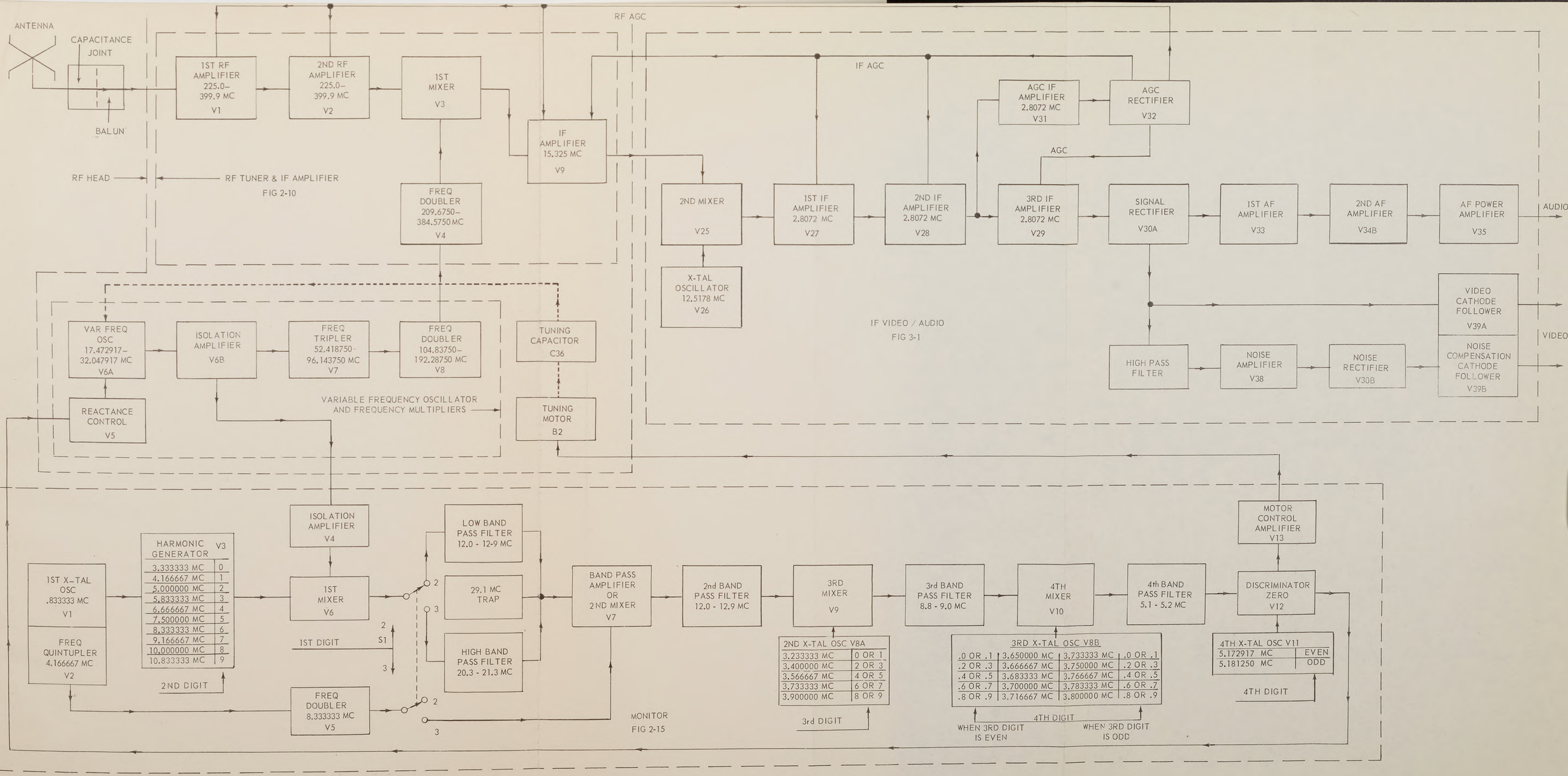
ANTENNA
CAPACITANCE JOINT
BALUN
1ST RF AMPLIFIER 225.0–399.9 MC V1
2ND RF AMPLIFIER 225.0–399.9 MC V2
1ST MIXER V3
IF AMPLIFIER 15.325 MC V9
RF AGC
IF AGC
AGC IF AMPLIFIER 2.8072 MC V31
AGC RECTIFIER V32
AGC
RF HEAD
RF TUNER & IF AMPLIFIER
FIG 2-10
FREQ DOUBLER 209.6750–384.5750 MC V4
2ND MIXER V25
1ST IF AMPLIFIER 2.8072 MC V27
2ND IF AMPLIFIER 2.8072 MC V28
3RD IF AMPLIFIER 2.8072 MC V29
SIGNAL RECTIFIER V30A
1ST AF AMPLIFIER V33
2ND AF AMPLIFIER V34B
AF POWER AMPLIFIER V35
AUDIO
X-TAL OSCILLATOR 12.5178 MC V26
IF VIDEO / AUDIO
FIG 3-1
VIDEO CATHODE FOLLOWER V39A
VIDEO
NOISE COMPENSATION CATHODE FOLLOWER V39B
HIGH PASS FILTER
NOISE AMPLIFIER V38
NOISE RECTIFIER V30B
VAR FREQ OSC 17.472917–32.047917 MC V6A
ISOLATION AMPLIFIER V6B
FREQ TRIPLER 52.418750–96.143750 MC V7
FREQ DOUBLER 104.83750–192.28750 MC V8
TUNING CAPACITOR C36
REACTANCE CONTROL V5
VARIABLE FREQUENCY OSCILLATOR AND FREQUENCY MULTIPLIERS
TUNING MOTOR B2
ISOLATION AMPLIFIER V4
LOW BAND PASS FILTER 12.0 - 12-9 MC
29.1 MC TRAP
HIGH BAND PASS FILTER 20.3 - 21.3 MC
1ST X-TAL OSC .833333 MC V1
FREQ QUINTUPLER 4.166667 MC V2
HARMONIC GENERATOR V3
3.333333 MC 0
4.166667 MC 1
5.000000 MC 2
5.833333 MC 3
6.666667 MC 4
7.500000 MC 5
8.333333 MC 6
9.166667 MC 7
10.000000 MC 8
10.833333 MC 9
2ND DIGIT
1ST MIXER V6
1ST DIGIT
S1
FREQ DOUBLER 8.333333 MC V5
BAND PASS AMPLIFIER OR 2ND MIXER V7
MONITOR
FIG 2-15
2nd BAND PASS FILTER 12.0 - 12.9 MC
3RD MIXER V9
2ND X-TAL OSC V8A
3.233333 MC 0 OR 1
3.400000 MC 2 OR 3
3.566667 MC 4 OR 5
3.733333 MC 6 OR 7
3.900000 MC 8 OR 9
3rd DIGIT
3rd BAND PASS FILTER 8.8 - 9.0 MC
4TH MIXER V10
3RD X-TAL OSC V8B
.0 OR .1 | 3.650000 MC | 3.733333 MC | .0 OR .1
.2 OR .3 | 3.666667 MC | 3.750000 MC | .2 OR .3
.4 OR .5 | 3.683333 MC | 3.766667 MC | .4 OR .5
.6 OR .7 | 3.700000 MC | 3.783333 MC | .6 OR .7
.8 OR .9 | 3.716667 MC | 3.800000 MC | .8 OR .9
4TH DIGIT
WHEN 3RD DIGIT IS EVEN
WHEN 3RD DIGIT IS ODD
4th BAND PASS FILTER 5.1 - 5.2 MC
MOTOR CONTROL AMPLIFIER V13
DISCRIMINATOR ZERO V12
4TH X-TAL OSC V11
5.172917 MC EVEN
5.181250 MC ODD
4TH DIGIT

Figure 2-4. –Block diagram of r-f head monitor, and video-audio chassis.

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to the vfo by operating a tuning capacitor (which is controlled by a tuning motor driven by the monitor) and by controlling a reactance tube shunted across the vfo.

Essentially, the frequency control system performs in the following manner: A sample of the output of the variable frequency oscillator in the r-f head is beat against a series of crystal-controlled frequencies in the monitor to produce a frequency to be fed to a crystal-controlled phase discriminator. The discriminator output, in turn, causes the frequency of the vfo to approach the correct frequency by controlling either the tuning motor or the reactance tube. The frequency at which the monitor operates is determined by the selection of the various crystal-controlled frequencies in the monitor.

An isolation amplifier stage in the r-f head couples the output of the vfo through another isolation amplifier to a first mixer circuit in the monitor. The first mixer receives another input from a harmonic generator, which, in turn, is supplied from the first crystal oscillator. This oscillator produces the fundamental frequency generated in the monitor circuit.

The harmonic generator produces frequencies ranging from the 4th to the 13th harmonic of the oscillator fundamental frequency. Either of these harmonics may be selected from the generator circuit by tuning a transformer (not shown). The selected output is heterodyned with the vfo frequency output in the 1st mixer.

If the monitor is set to a frequency in the 225- to 299.9-mc range, a selector switch permits the output of the 1st mixer to be fed to a low band-pass filter. This signal is subsequently passed to a band-pass amplifier, and to a 2nd band-pass filter. When the monitor is set to a frequency in the 300.0- to 399.9-mc range, the 1st mixer output is fed to a high band-pass filter and, in turn, is mixed with the tenth harmonic of the 1st crystal oscillator at the 2nd mixer. This action produces a beat frequency, which is fed to a 2nd band-pass filter. The band-pass amplifier or 2nd mixer stage functions as an amplifier for frequencies from 225 to 299.9 mc and as a mixer stage for frequencies from 300.0 to 399.9 mc. The tenth harmonic of the 1st crystal-oscillator frequency is produced by the action of a frequency quintupler and frequency doubler, which are cascade connected between the oscillator and 2nd mixer circuits.

A 2nd crystal oscillator provides a choice of five frequencies. The 3rd mixer combines the output of the second band-pass filter with the selected output of the 2nd crystal oscillator to produce a beat frequency. This output is fed through the 3rd band-pass filter to the 4th mixer. The 3rd crystal oscillator provides a choice of 10 frequencies. Either of five of these frequencies may be selected when the chosen frequency has an even third digit, with either of the other five frequencies being selected when the third digit of the chosen frequency is an odd number. The selected crystal-controlled frequency is heterodyned in the 4th mixer stage with the input from the 3rd band-pass filter. The resulting output is passed through the 4th band-pass amplifier to the discriminator. The frequency applied to the discriminator will always be 5.172917 mc when the fourth digit is even, and 5.181250 mc when the fourth digit is an odd number.

In order to gain a better understanding of the monitor operation, consider the following example.

Assume that a frequency of 323.0 mc is selected by the channel selector switch (fig. 2-1) as the input frequency to the r-f head. The variable frequency oscillator (fig. 2-4) tracks the r-f input, and produces an output, which will be multiplied 12 times in the frequency multiplier stages (3 x 2 x 2), and applied to the 1st mixer in the r-f head, yielding a frequency 15.325 mc lower than the incoming r-f (323.0 mc). The 15.325-mc difference frequency can therefore be subtracted from the incoming r-f signal to yield the frequency being applied from the multiplier stages to the 1st mixer in the r-f head. Thus,

$$\begin{array}{r} 323.000000 \text{ mc} \\ -\ 15.325000 \text{ mc} \\ \hline 307.675000 \text{ mc} \end{array}$$

If the 307.675-mc frequency is divided by 12, the variable frequency oscillator output frequency will be derived.

$$12\overline{)307.675000} \quad 25.639583 \text{ mc}$$

Thus, the variable frequency oscillator operates at 25.639583 mc when the r-f input (at the antenna) is 323.0 mc. The monitor output must therefore control the vfo at this frequency, provided the input frequency remains 323.0 mc.

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A portion of the vfo output frequency (25.639583 mc) is fed through two isolation amplifiers to the 1st mixer in the monitor. The second digit of the incoming frequency (323.0 mc) determines which of the harmonics from the harmonic generator will be heterodyned with the vfo input in the 1st mixer. Because the second digit of the assumed frequency is 2, the 5.0-mc output of the harmonic generator will be coupled to the 1st mixer. The resulting difference frequency output of the 1st mixer will be

$$\begin{array}{r} 25.639583 \\ -\ 5.000000 \\ \hline 20.639583 \text{ mc} \end{array}$$

When the frequency selected by the channel selector (fig. 2-1) is in the range from 225.0 to 299.9 mc, the switch, S1, at the output of the 1st mixer in the monitor (fig. 2-4) will be in the position shown (up). For input frequencies from 300.0 to 399.9 mc, S1 is in the position opposite to that shown (down).

Since the chosen frequency is 323.0 mc, S1 at the 1st mixer output in the monitor is down (opposite to that shown), and the output frequency of the mixer is fed through the high band-pass filter to the 2nd mixer. (The band-pass amplifier or 2nd mixer operates as a mixer stage when the input frequency is above 300.0 mc, as discussed.)

With S1 in the down position, the 10th harmonic of the 1st crystal oscillator in the monitor (8.333333 mc) is heterodyned with the high band-pass filter input in the 2nd mixer. The difference frequency produced is

$$\begin{array}{r} 20.639583 \\ -\ 8.333333 \\ \hline 12.306250 \end{array}$$

The 2nd mixer output is fed through a 12.0- to 12.9-mc, 2nd band-pass filter to the 3rd mixer. A second input, from a 2nd crystal oscillator, is also applied to the 3rd mixer. The selected frequency from the 2nd crystal oscillator is dependent upon the value of the third digit in this input frequency. For a frequency, 323.0 mc, the third digit is 3. Likewise, the selected frequency to be applied to the 3rd mixer will be 3.400000 mc. Thus, the difference frequency output of the 3rd mixer is

$$\begin{array}{r} 12.306250 \\ -\ 3.400000 \\ \hline 8.906250 \text{ mc} \end{array}$$

The 8.906250-mc output of the 3rd mixer is fed through an 8.8- to 9.0-mc, 3rd band-pass filter to the 4th mixer. If this third digit of the incoming frequency is an even number, the third crystal oscillator provides either of the five frequencies indicated on the left. If the third digit of the incoming frequency is an odd number, either of the five frequencies indicated on the right is selected. The frequency selected from the left or right group is dependent upon the value of the fourth digit. The frequency selected for a 323.0-mc input is 3.733333 mc. Thus, the input frequency to the discriminator will be

$$\begin{array}{r} 8.906250 \\ -3.733333 \\ \hline 5.172917 \text{ mc.} \end{array}$$

Since the fourth digit of the input frequency (323.0 mc) is even, the 4th crystal oscillator also supplies an input of 5.172917 mc to the discriminator.

The discriminator is a phase-sensitive device, which develops an output voltage representing the resultant of any phase deviation between the two frequencies. When the two frequencies are exactly 90° out of phase, the discriminator error voltage produced represents zero error. Any deviation in phase develops either a positive or negative discriminator voltage output, the polarity depending upon whether the 4th band-pass filter frequency is leading or lagging the 4th crystal-oscillator frequency in phase. The amplitude is dependent upon the phase angle existing between the two frequencies.

If the error voltage is large, the discriminator output voltage is fed to a motor control amplifier whose output controls a tuning motor. The tuning motor, in turn, drives a vfo tuning capacitor. The discriminator output and associated circuits cause the tuning motor to tune the vfo to that frequency which, when fed through the monitor unit, is approximately the same frequency as that of the 4th crystal oscillator.

When the frequency is close to that of the desired operating frequency, the reactance tube in the r-f head takes control of the vfo to bring the circuit to the exact frequency. This action is described later. Since the discriminator voltage output tunes the vfo to a frequency that equals the sum of all the crystal-controlled frequencies, it follows that the vfo frequency may be changed a given amount by changing any one of the crystal-controlled frequencies the same amount.

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The output of the vfo is coupled through an isolation amplifier and multiplier stages to the 1st mixer in the r-f head. The 1st and 2nd r-f amplifiers in the r-f head amplify the input signal to a level sufficient to provide proper heterodyning of the two signals. The 1st mixer output of 15.325 mc is amplified in the 1st i-f stage and coupled through a coaxial cable to the i-f, video-audio chassis. The r-f and 1st i-f amplifiers are AGC controlled.

A second frequency heterodyning process takes place at the input to the i-f, video-audio chassis to produce a 2.8072-mc, i-f frequency. The remaining circuits of the i-f video-audio chassis are conventional receiving circuits.

Due to distance limitations of the receiver from the antenna, the indicator unit is normally situated at a remote location relative to the receiving and antenna equipments. The frequency of operation of the equipment is usually selected remotely from the front panel of the indicator.

During channel selection, the setting of tabs on a "memory" cylinder (fig. 2-5) determines which 20 of 1750 available frequencies will be selected by remote control as the (preset)

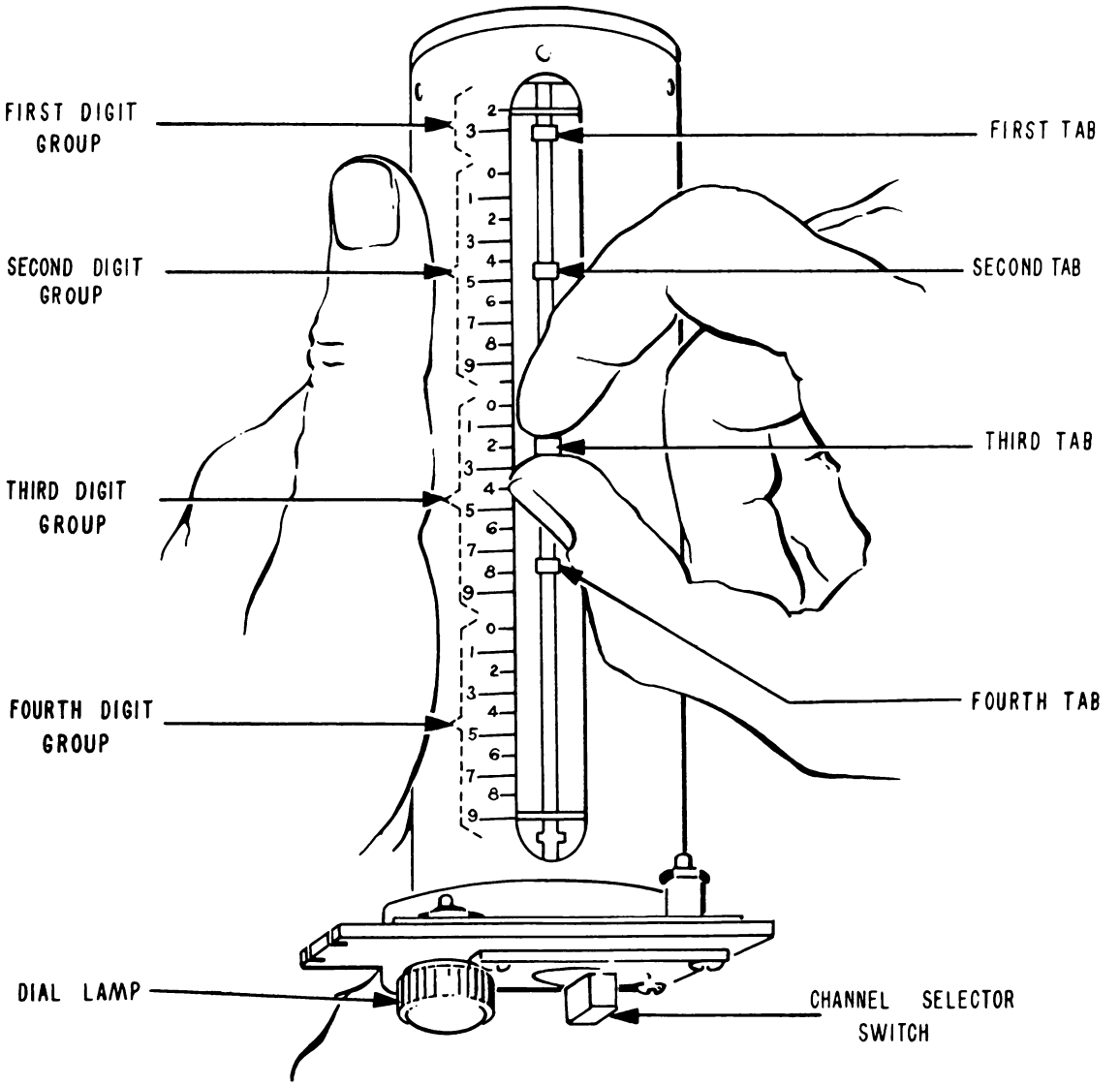

Figure 2-5. –Memory cylinder adjustment.

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channel frequencies. When the preset channel selector switch (fig. 2-1) is turned from one position to another, bridge circuits connected between the monitor and indicator are unbalanced, and an automatic selection of a new frequency is begun. The unbalanced circuits cause the operation of relays, which rebalance the circuits for the new positions of the channel selector switches. The positions of these switches (which balance the bridge circuits) are also the switch positions, which select the proper combination of frequencies to be fed to the harmonic generator and the 2nd, 3rd, and 4th crystal oscillators in the monitor circuit.

Mounted just behind the front panel of the indicator is a 20-position rotary turret assembly (fig. 2-5). The memory cylinder fits inside the removable turret assembly and mechanically actuates various contacts of the channel selector switch. Arranged around the periphery of the memory cylinder are 20 evenly spaced detented bars, which are parallel to its longitudinal axis. Along each bar are four tabs, which can be positioned laterally along the length of the cylinder. The bars and their tabs are accessible through a slot in the cylinder case. By positioning the four tabs on a particular bar in relationship to a series of indexing digits on the cylinder, a particular channel frequency can be set up. Since there are 20 longitudinal bars along the cylinder, it follows that 20 of the total 1750 frequencies available within the tuning range of the equipment can be selected as channel operating frequencies by presetting the cylinder.

A block diagram of the indicator unit is shown in figure 2-6. The indicator provides the means by which the detected signal is amplified and mixed with the sweep voltages, which were originated at the antenna. The indicator also presents an oscilloscopic display of the received direction finding information.

The alternator voltages from the antenna are fed to the indicator, as discussed. These voltages serve as the horizontal and vertical sweep input voltages, respectively. The phase of the vertical input is shifted 90° from that of the horizontal to produce a circular (Lissajous) pattern on the cathode-ray indicator in the absence of a video signal.

The input video signal is amplified and fed through a sense D/F calibrate switch to the first modulator circuit when the switch is in the sense D/F position. A 200-kc carrier oscillator output is also fed to the first modulator, and is therefore modulated in this stage by the video signal. The resulting signal output is amplified and applied from a cathode follower circuit to both the vertical and horizontal modulator stages.

The input video signal is applied to the video differential amplifier. The video differential amplifier is a duo-triode tube, one section of which receives a d-c control voltage whose magnitude is proportional to the amplitude of the prevailing noise. The other section of the tube receives a negative-going video signal. The video output of the video section causes the current in the noise section of the video differential amplifier to vary.

When the video input is maximum negative (representing the maximum input at the antenna) the current through the noise section of the tube is maximum, and is effective in supplying a negative-going grid voltage, which cuts off the first modulator tube. When the first modulator is cut off, no video signal can pass through the subsequent stages to the indicator, and a spot appears in the center of the indicator screen.

Conversely, when the video input is minimum negative, the conduction of the noise section of the video differential amplifier is minimum, and the first modulator stage is effective in supplying a video signal through the subsequent stages to the indicator screen. Thus, a video signal is fed to the indicator during the time of the nulls at the antenna, and the tips of the propeller-shaped pattern produced on the indicator points along the nulls of the received signal.

If the sense D/F switch is placed in the CALIBRATE position, the calibration voltage (also originated at the antenna) is fed to the modulator. Thus, in this condition, the calibration voltage modulates the 200-kc oscillator carrier, and is subsequently applied to the horizontal and vertical modulators.

The calibration voltage input is derived from a 36-toothed tone wheel (see figure 2-8), which generates 36 cycles of input voltage for each rotation of the antenna shaft. The calibration input thus operates the first modulator 36 times as often as the video signal, and the pattern consisting of 36 propeller-shaped lobes is presented on the indicator for calibration purposes.

The chosen modulated voltage applied to the vertical and horizontal modulators is further modulated by the alternator voltage input (from the antenna) to each of the modulator stages.

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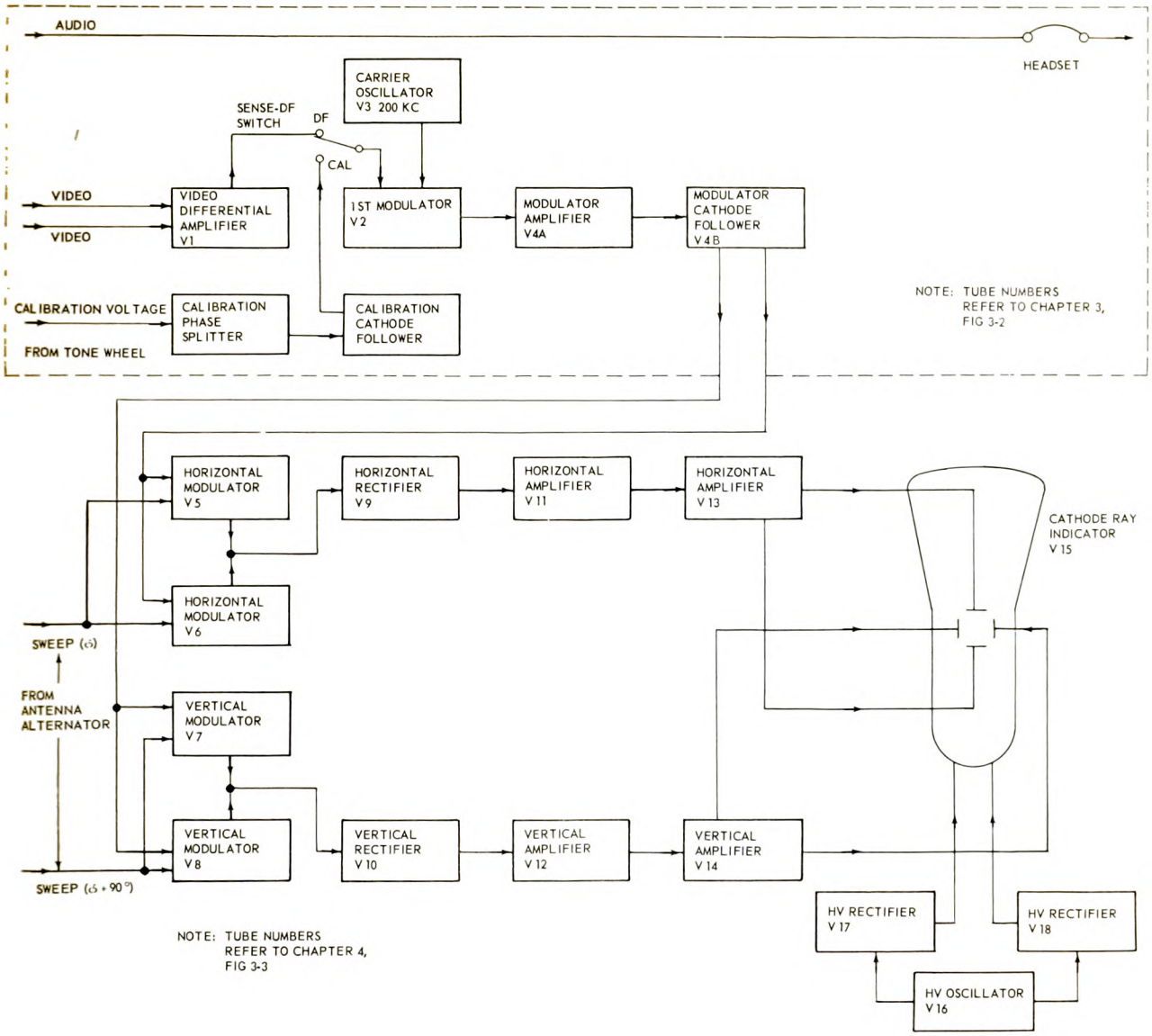

Figure 2-6. –Azimuth indicator, block diagram.

The resultant voltage from the separate horizontal and vertical modulators is rectified, amplified (in two stages) and subsequently applied to the vertical and horizontal deflection plates, respectively. The modulation process in the horizontal and vertical modulators, in the presence of a video signal, produces a voltage, which when applied to the respective deflection plates causes the pattern to assume a propeller shape (fig. 2-3, A).

A sense circuit is put into operation when it is desired to determine the direction from which the signal is being received. Normally, both sides of the antenna are in exact balance. The sense circuit causes the operation of a relay (shown later), which alters the electrical characteristics of the antenna. The electrical change of the antenna creates an unbalance, and a phase shift in the field response pattern, which causes the propeller-shaped pattern on the cathode-ray indicator to be distorted. The distorted pattern is in the form of a distorted figure 8 (fig. 2-3, B), the apex of which points in the direction of the signal origin.

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As stated earlier, when the sense D/F calibrate switch is in the sense D/F position, the video signal will be displayed as a propeller-shaped pattern. When the switch is in the CALIBRATE position, calibration voltage will be displayed as a 36-lobe pattern (fig. 2-7), which provides a means of calibrating the indicator.

The high voltage for the cathode-ray tube is obtained by rectifying an r-f voltage generated by an r-f oscillator. The voltage is applied to the accelerating anode of the tube.

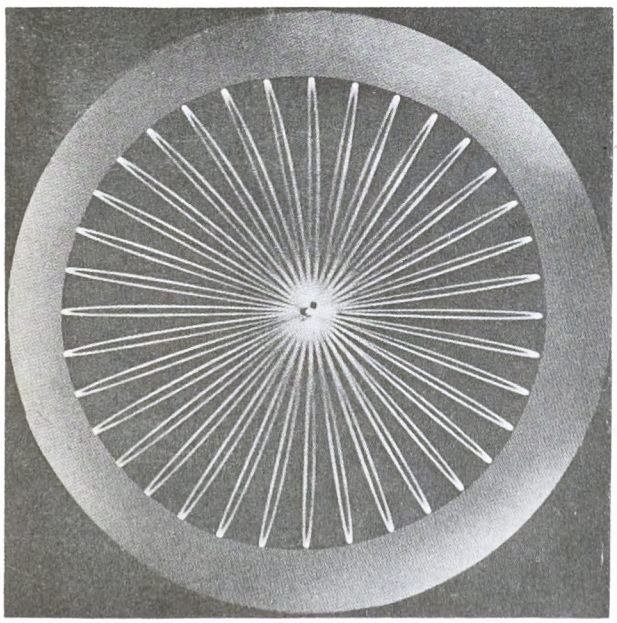

Figure 2-7. –Indicator calibration pattern.

RADIO RECEIVER

ANTENNA

The antenna (fig. 2-8) is a two-dipole Adcock array, which is rotated by a direct-coupled, 900-rpm induction motor, B1. The motor also drives an alternator, G1, and a wheel, E1, for purposes that are discussed later.

Normally, the antenna is electrically connected, as represented in figure 2-9, A. Maximum signal voltage will be induced into both antennas simultaneously when the signal is received, as shown in the illustration. Since the dipoles are one-half wavelength at the mean frequency, the voltage induced in the two halves of each dipole will be 180° out of phase. Likewise, the separate dipoles are spaced one-half wavelength apart (at the mean frequency), causing the voltage in the two dipoles to be out of phase 180°.

Assume that the received signal induces a voltage in the antenna of the polarity shown in figure 2-9, A. An antenna current is conducted for dipole 1, which is through L2, C2, T1, C1, and L1. Dipole 2 conducts through L4, C2, T1, C1, and L3. Note that the two signals reinforce each other at the balun. Thus, the receiver input is approximately twice that received by a single antenna.

Relay K1 (fig. 2-8) alters the circuit of the antenna when the sense D/F calibrate switch on the panel of the indicator (fig. 2-1) is moved to the SENSE position. The circuit then appears as shown in figure 2-9, B. In this condition, a phase-shifting network comprising L1, L2, R1, and R2 is connected across the dipole 1 input. Thus, the phase difference between the two voltages from both dipoles is shifted to alter the pattern of the indicator. In addition, R3 and R4 are shunted across L3 and L4, respectively, causing attenuation of the dipole 2 input. The effect of the phases to produce the propeller-shaped indicator pattern and the "sense" (distorted figure 8) display on the cathode-ray indicator will be considered later. Power for operating K1 is obtained from terminal 12 of TB1 (fig. 2-8) via the sense D/F calibrate switch (see chapter 3, figure 3-3).

Due to antenna rotation (approximately 900 rpm) the r-f voltage induced into each dipole (fig. 2-8) from a given source is caused to vary at a sinusoidal rate. Thus, the input to the balun (and subsequently to the receiver) is a modulated r-f voltage.

The output of the antenna array at the antenna terminals is balanced to ground. To permit the use of a coaxial cable for transferring the output of the array to the input of the receiver, the antenna output is made unbalanced. The balance-to-unbalance condition is created by a balun, T1.

The rotating antenna array is coupled to the stationary portion of the antenna through rotating capacitances C3 and C4 (fig. 2-8). The energy is fed into the balun one-quarter wavelength above ground. The voltage at this point is maximum. A coaxial cable connected between "c" and "d" of the balun supplies the input to the receiver.

The antenna drive motor, B1 (fig. 2-8) is an induction-type, single-phase motor, which drives a single antenna shaft at 900 rpm (15

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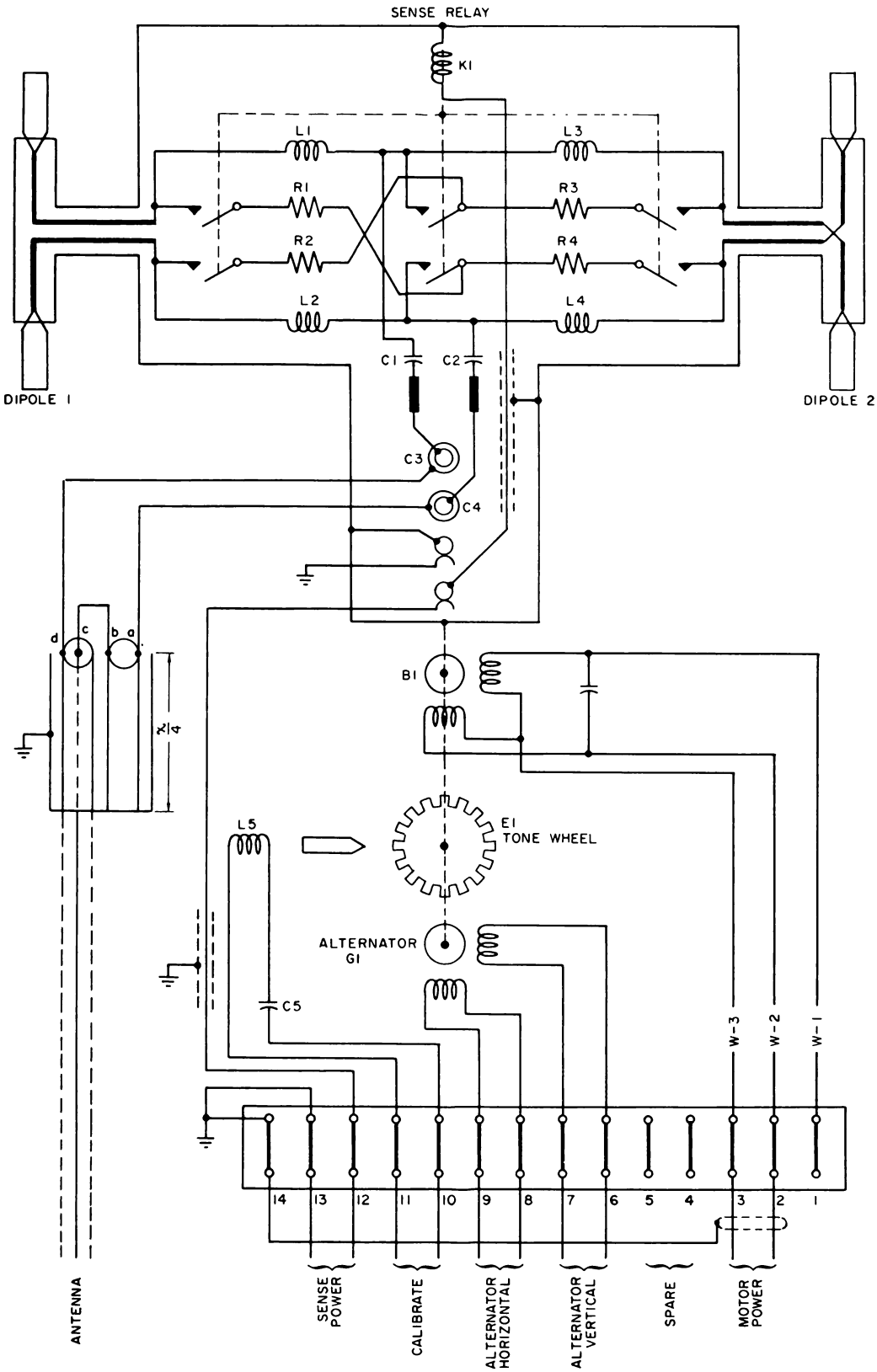

Figure 2-8.—Antenna schematic diagram.

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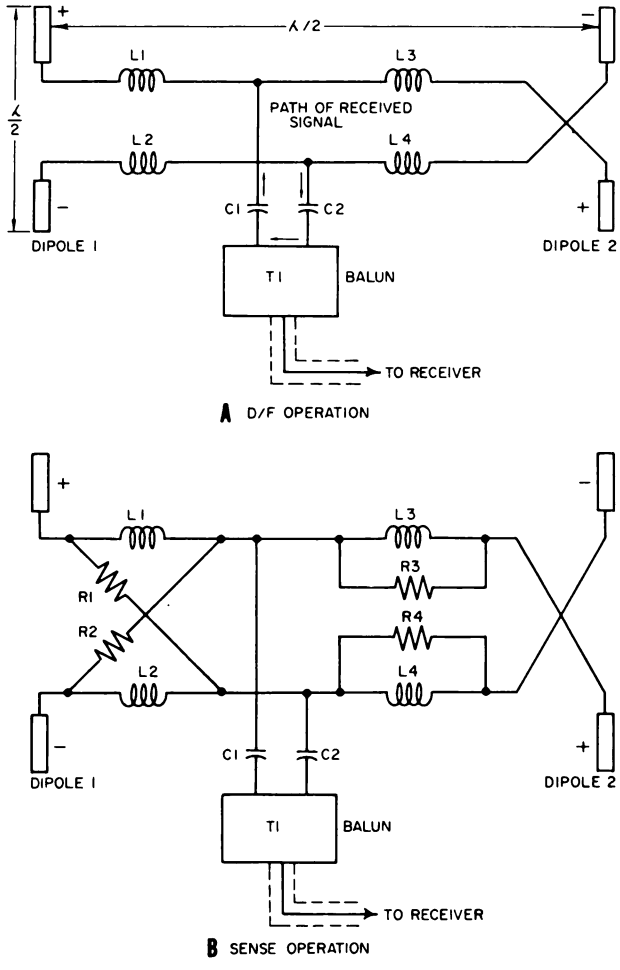

Figure 2-9. –Antenna, schematic for D/F and sense operation.

rps). The motor also drives a two-phase alternator, G1. The alternator produces one cycle of sweep voltage for the horizontal and vertical deflection plates, respectively, for each complete rotation of the antenna. Thus, the frequency of the sweep voltage to the cathode-ray tube deflection plates is 15 cps.

A 36-lobe calibration pattern (fig. 2-7) is presented on the indicator tube when the sense D/F calibrate switch (fig. 2-1) is placed in the CALIBRATE position. One lobe appears for each 10 degrees of the 360-degree total.

A 36-toothed tone wheel, E1, driven by the antenna drive motor, affects the highly permeable core of L5 (fig. 2-8) to induce an alternating current into the circuit comprising L5, C5, and the calibration input circuit (connected between 10 and 11 of TB1). C5 tunes L5 so that the voltage developed is correct to cause the production of the 36-lobe pattern. The manner by which the induced voltage affects the indicator circuits is treated later.

R-F HEAD

The signal from the antenna is fed via C1 (fig. 2-8) to the r-f head of the receiver (fig. 2-10). The input is a 15-cps modulated r-f voltage in the 225.0- to 399.0-mc range.

The first r-f amplifier, V1, the second r-f amplifier, V2, the mixer, V3, and the frequency doubler, V4, are capacitance tuned by capacitors contained in their respective cavities. This type of tuning is utilized to obtain a high Q and to provide frequency stability.

To understand the operation of the cavities, consider the diagram (fig. 2-11, A), which illustrates the physical construction of the cavity. Figure 2-11, B, shows the electrical equivalent of the circuit. The center post of the cavity acts as an inductance, L1. In conjunction with ganged capacitors C1, C2, and C3, L1 is electrically equivalent to a quarter-wavelength rod whose base is grounded, and whose opposite end is at a point of high impedance. The cavity is tuned to resonance by C1, C2, and C3. C4 is a trimmer capacitor, and slug Lv is a tuning vane, which can be utilized to adjust the inductance, L1, of the cavity.

The r-f input via J1 (fig. 2-10) is tapped down on the inductance of the cavity at a point that matches the impedance of the cable carrying the antenna signal to the input impedance of the r-f amplifier, V1. The first and second r-f amplifiers are connected as grounded grid circuits to lessen the possibility of self-oscillations in these stages by means of feedback through the tube capacitances. The grid of V1 is held at r-f ground potential by C49 and C50. An AGC voltage from the i-f video chassis (discussed later) is applied to the grids of V1 and V2, and thus controls the gain of these stages.

The input signal for the V1 and V2 amplifiers is taken from a point on their respective cavity inductances relative to ground. L9 is the plate load inductance of V1. The V1 output is coupled by C51 to a point of high impedance at the top of inductance L3. The operation of the two r-f amplifiers is identical.

The second r-f amplifier output is coupled to the grid of the mixer stage, V3, by L6 located in the mixer cavity. A second input is applied to the mixer stage from the frequency doubler,

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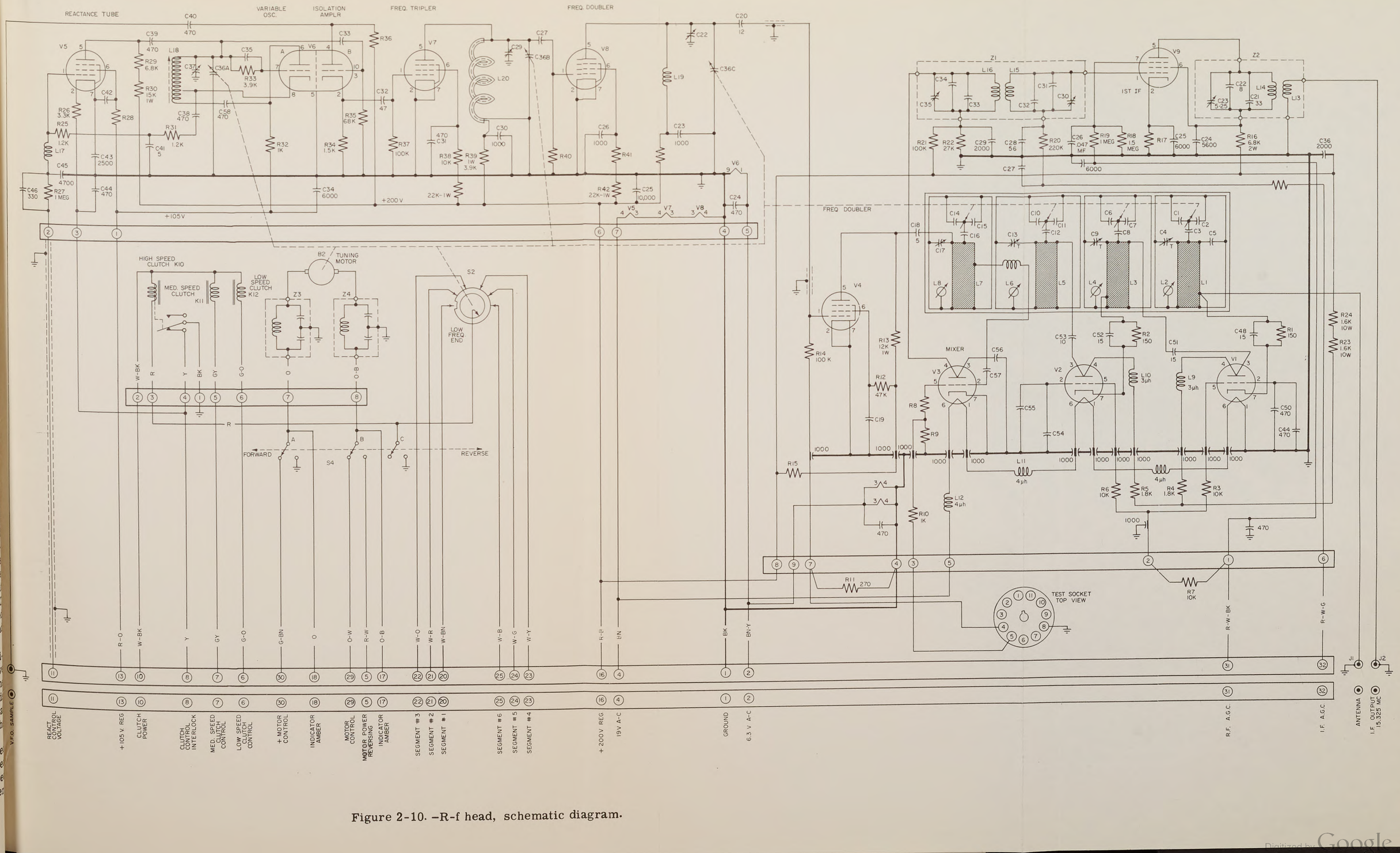

Figure 2-10. –R-f head, schematic diagram.

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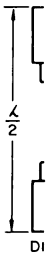

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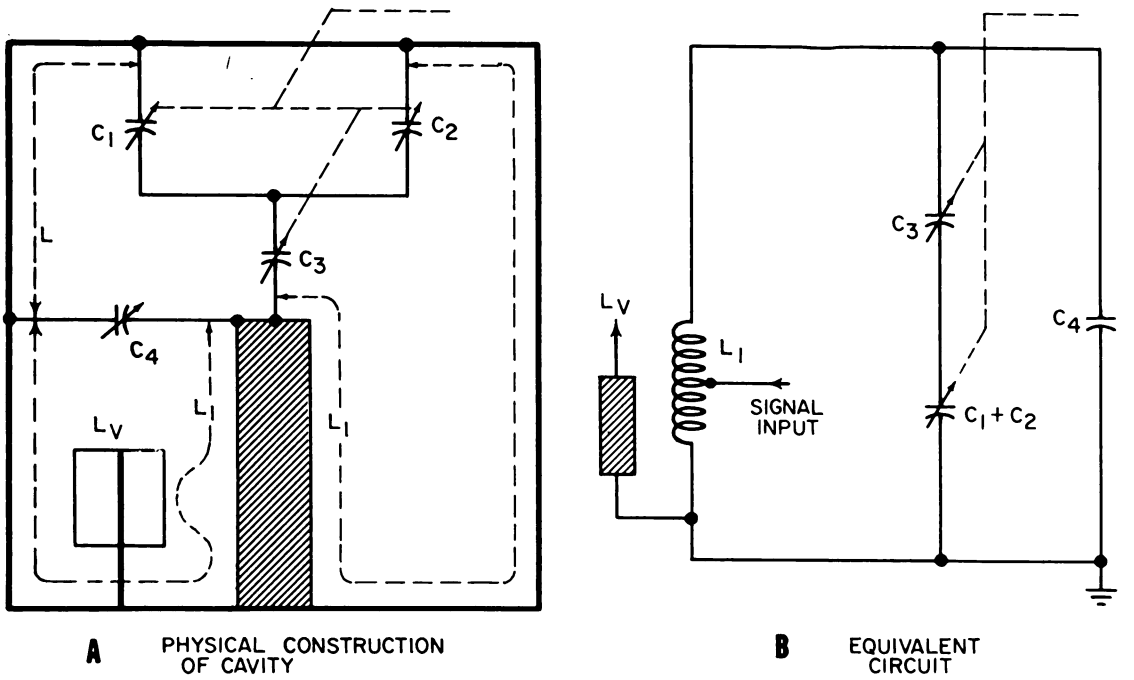

Figure 2-11. –R-f head cavity, physical and electrical equivalent circuits.

V4 (a part of the variable frequency oscillator). The input to V3 is taken from a point on L7 which develops the proper signal amplitude for heterodyning of the signals at the mixer. The mixing action occurs in V3 rather than the cavity since the tube's operation is non-linear and therefore capable of producing an output different than that of the input. The L7 tap also impedance matches the frequency doubler output circuit to the input circuit of the mixer.

VARIABLE FREQUENCY OSCILLATOR AND REACTANCE TUBE CIRCUITS

It should be recalled that the frequency of the mixer output is 15,325 mc. During frequency selection, reactance tube V5 receives a d-c control voltage from a discriminator circuit, which controls the frequency output of the variable frequency oscillator. The vfo circuit comprises tubes V5, V6, V7, V8, and V4, and produces the input frequency, which is fed to the V3 mixer cavity from the tap on L7.

The operation of the discriminator circuit to produce the d-c control voltage is treated in detail later. It is sufficient at this time to remain aware of the control voltage whose polarity may be either positive or negative, depending upon the direction of the error frequency from the desired frequency. In actual operation, zero error voltage from the discriminator is slightly positive to ground.

The reactance of C41 in series with C38 (in the reactance tube circuit (fig. 2-12, develops a capacitive reactance, which is equal to or greater than the resistance of R31. The r-f voltage from the tap of L18 to ground is also developed across C38, R31, and C41. That voltage across C41 is applied through R25 to the grid of the reactance tube, V5. Due to the capacitances involved, the r-f voltage at the grid of V5 lags the L18 (variable frequency oscillator) voltage by approximately 90°). The plate current of V5 is always in phase with the grid voltage. Thus, the plate r-f current lags the vfo oscillator r-f voltage by an angle that approaches 90 degrees. V5 is therefore an inductive reactance shunted across the oscillator circuit.

The d-c control voltage is fed through a decoupling network comprising C46 and R27 to the L17-C45 junction. Assume that this voltage goes more positive to ground. V5 will conduct a heavier current, and therefore acts as a smaller inductive reactance across the oscillator to increase the oscillator frequency output. Conversely, if the control voltage goes in a negative

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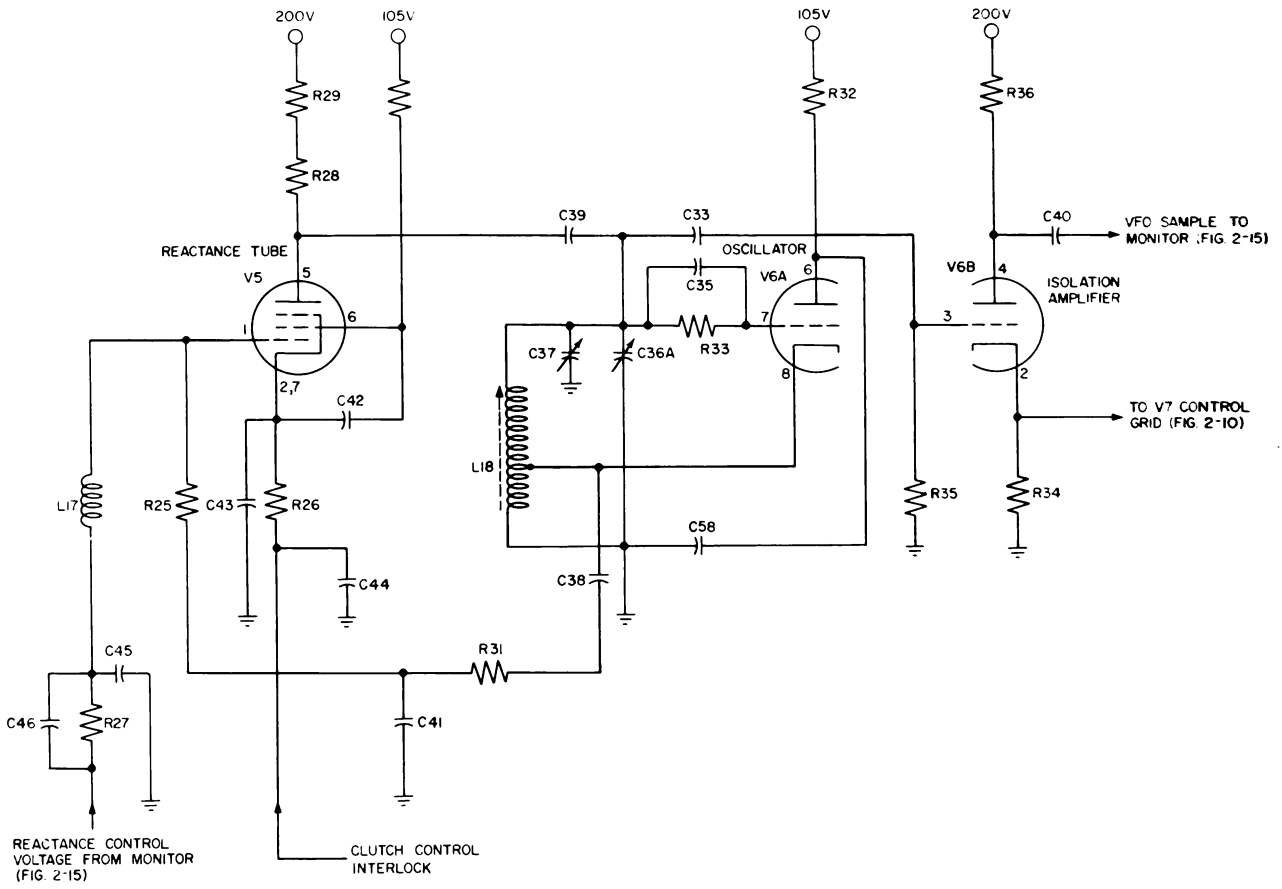

Figure 2-12. –Reactance control circuit.

direction, a larger lagging voltage is applied to the oscillator, and the oscillator frequency decreases. Thus, by controlling the bias at the grid of V5 with the d-c control voltage, the tube current and, in turn, the inductive reactance (as seen by the vfo) is also controlled. Cathode resistor R26 develops a bias of such a value that the control voltage up to approximately +5 volts, d-c is the equivalent of a zero error signal from the monitor.

The variable frequency oscillator circuit, V6A, is connected as a series-fed Hartley oscillator whose feedback is developed between the lower and upper sections of L18. C36A provides oscillator tuning, and C37 is the trimmer capacitor. The corrected oscillator output frequency is coupled by C33 to the control grid of a cathode follower, V6B, which serves as an isolation stage between the oscillator and frequency tripler stage, which follows.

The frequency tripler stage, V7 (fig. 2-10) receives its input from the cathode of V6B via C32. Grid leak bias is formed by C32 and R37. The positive portions of the grid signal drive the V7 grid to conduction to produce a negative-going plate output pulse for each input cycle. The plate tank comprising L20, C29, and C36B produces three output cycles for each plate pulse input, thereby tripling the frequency output as it is applied through C27 to the V8 grid. L20 is enclosed in copper tubing, which inductively couples the coil to C29 and C36B.

V8 acts as a frequency doubler stage. Doubling is accomplished in the plate tank circuit in a manner similar to that just discussed for the tripler stage. The V8 output is again doubled by frequency doubler stage V4.

The V4 frequency doubler output is applied to a high impedance point on the doubler tank inductance, L7. The frequency of this signal is 12 times that at the output of the vfo, V6A. With the vfo frequency corrected, the heterodyning process in the mixer stage, V3, produces

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a 15.325-mc signal, which is fed directly to a filter circuit, Z1.

Z1 comprises high Q, double-tuned, parallel tank circuits whose peak response occurs at 15.3 mc. The output of Z1 is coupled to the grid of the first i-f amplifier, V9. This stage, like the first and second r-f amplifiers, is AGC controlled. V9 is a conventional i-f amplifier stage whose output is fed, via J2, to the i-f video-audio chassis.

FREQUENCY SELECTION CONTROL CIRCUITS

As stated earlier, the memory cylinder located behind the front panel of the indicator can be preset to select any one of 20 of the total 1750 available frequencies. Frequency selection is initiated by the four lateral bars (fig. 2-5), which mechanically actuate various contacts of the channel selector switch.

When one of the preset frequencies is to be selected, the channel selector switch is turned to the new position, which causes the unbalance of bridge circuits. These circuits are similar to that shown in figure 2-13. Three similar bridge circuits exist, corresponding to the second-, third-, and fourth-digit controls of the frequency selection system.

To select the new frequency, the channel selector switch chooses a tap on its associated bridge resistor, which corresponds to the position of the actuating bar on the memory cylinder. It should be recalled from the block diagram discussion that the monitor circuits produce a discriminator d-c control voltage, which controls the frequency output of the vfo in the r-f head. Thus, a change in the input frequency to the r-f head necessitates a change in the monitor frequencies so that the vfo will be locked-in at the correct frequency.

The channel selecting bridge circuits (fig. 2-13) are established between the indicator and monitor. If that section of the channel selector switch (for a particular bridge) and an associated rotary selector switch in the monitor are not at the same tap on their respective resistors, a difference in potential will exist at the ends of the bridge. The voltage thus established is placed across a bridge relay, K4, and the relay is energized.

With K4 energized, its contacts close to connect a clutch relay, K1 (fig. 2-14) between a +27.5-volt-supply and ground. The 3-4 contacts of K1 close to apply motor power to the selector

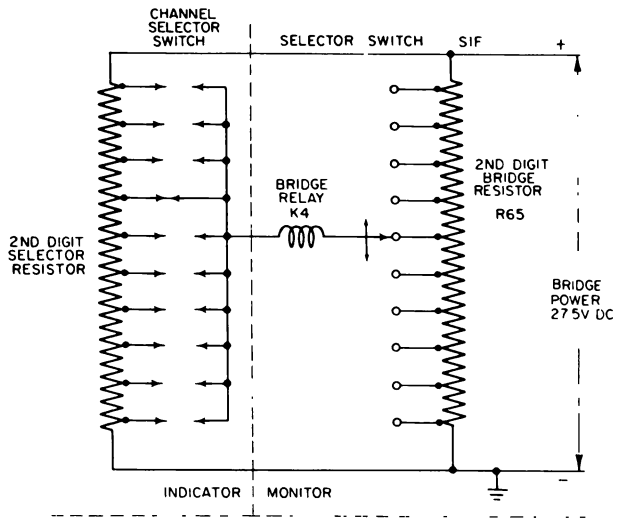

Figure 2-13. —Bridge tuning circuit diagram.

motor, B1. The clutch relay also releases its associated clutch face by pulling a pawl, P, out of a notch on a detent plate, CP1. The released clutch face allows the selector motor to operate, driving the wiper on the rotary selector switch S1F, until it reaches a tap, which again balances the bridge. The bridge and clutch relays are then deenergized, and the clutch pawl drops into a slot on the detent plate, which stops the arm of the rotary selector switch.

When the motor becomes energized, it continues to rotate until each of the three bridge circuits are balanced. Clutches belonging to selector switch sections that reach their positions continue to slip (as a result of motor drive) until all bridge circuits are again balanced. When the last bridge relay deenergizes, the clutch relay also deenergizes, and the motor stops.

Sections of the selector switch are ganged to their respective clutch plates (fig. 2-15). The clutch plates CP1, CP2, and CP3 (when released) drive their respective switch sections to select the proper combination of components in the harmonic generator and mixer, and the second, third, and fourth crystal oscillators of the monitor. The action is treated further in the discussion of the monitor.

MONITOR

The monitor (fig. 2-15) functions to control the frequency output of the variable frequency

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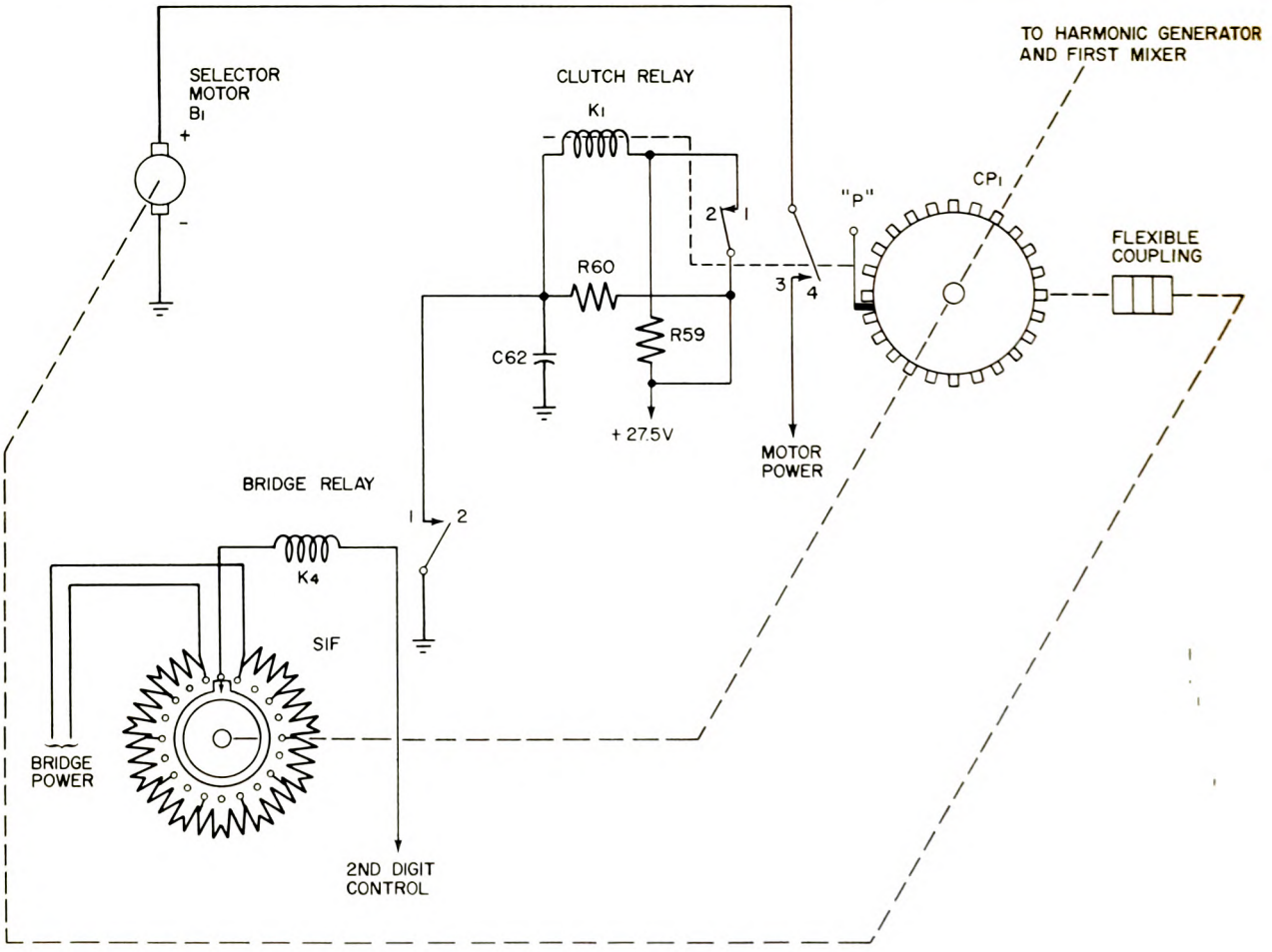

Figure 2-14. –Frequency selection control circuits.

oscillator in the r-f head. This action is accomplished by controlling the tuning capacitor, C36, of the vfo (fig. 2-10) by means of a motor driven by the monitor, and by controlling the reactive properties of a reactance tube shunted across the vfo.

Stage V2 (fig. 2-15) serves as the crystal-controlled, first crystal oscillator, which generates a fundamental frequency of 0.833333 mc. The grid of the stage is grounded to decrease the effects of signal frequency shunting due to the capacitance of the tube. The temperature of the crystal is maintained constant in a crystal oven. The crystal voltage is obtained from the difference in potential of the cathodes of V1 and V2. The crystal input signal developed at the V2 cathode is amplified and fed back to the crystal by the conduction of V1 through R4. This feedback signal is regenerative, and thereby tends to sustain the oscillations of V2.

V1 develops two outputs. One output is taken from the V1 cathode and used for feedback purposes, as discussed. The second output is multiplied by five in the plate circuit by L1 and C2, and coupled by C1 to the grid of the frequency doubler stage, V5. The purpose of this signal will be considered later.

A portion of the V2 output (fig. 2-15) is developed across the tank circuit comprising C7 and L2. This output is coupled by C9 to the grid of the harmonic generator, V3. The plate circuit of V3 contains a low Q tank circuit comprising the T1 primary, R12, C21, C20, and any one of ten capacitors that may be shunted across the tank by S1B. The 10 capacitors, C10 through C19 (in conjunction with C20 and C21) make possible the generation of any harmonic

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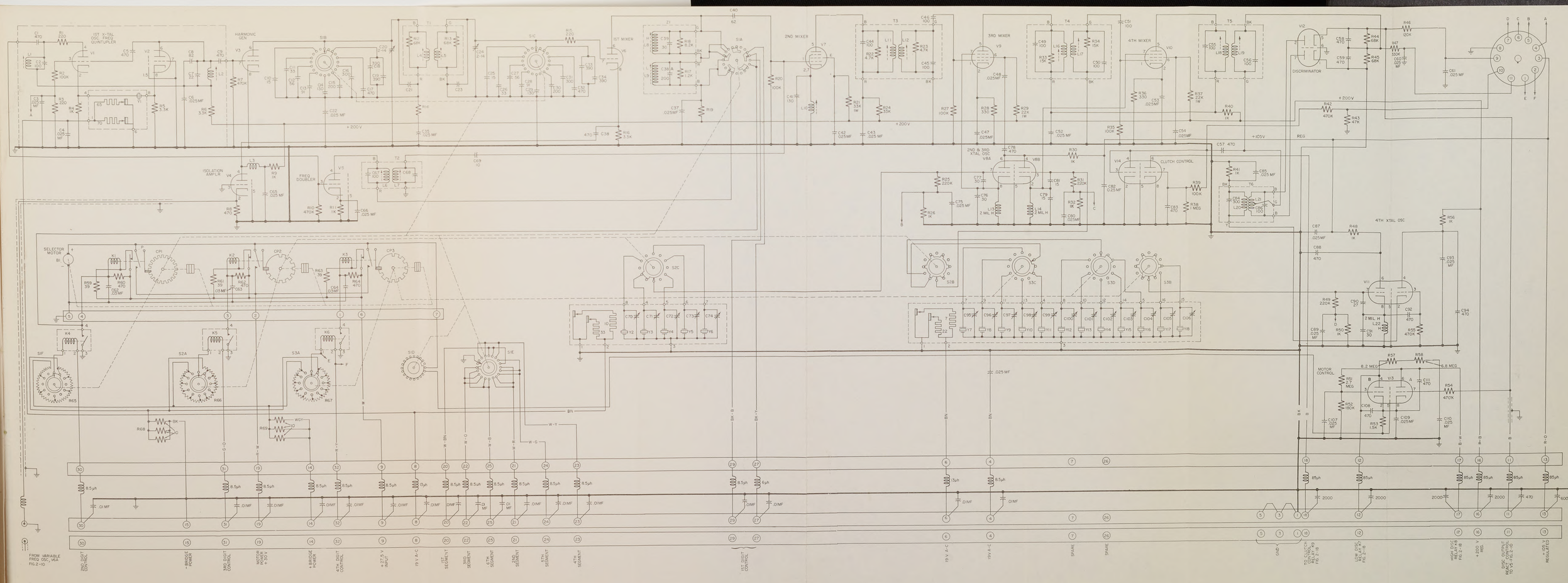

Figure 2-15. –Monitor, schematic diagram.

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of the fundamental frequency from the fourth to the 13th harmonic.

During the period that the selector motor is active, clutch plate CP1 drives the A, B, and C sections of S1. When the motor rotation ceases, the primary and secondary of T1 will be shunted by the proper value of capacitance (selected by S1B and S1C) to tune the T1 input and output to the desired frequency. S1A chooses either of two filters. L8, C39, and R18 form a high-frequency band filter circuit, and L9, C38, and R17 form a low-frequency band filter.

It should be noted from the block diagram (fig. 2-4) that the heterodyning process in the monitor should always produce a frequency at the discriminator input of either 5.172917 or 5.181250 mc.

The first mixer stage, V6 (fig. 2-15) receives the harmonic generator output at its grid while the V6A vfo output (fig. 2-10) is applied through a grounded grid amplifier, V4 (fig. 2-15) to the V6 cathode through C38. Thus, the two frequencies are heterodyned in the first mixer stage. If the equipment is set to receive a frequency in the 300.0- to 399.9-mc range, S1A selects the high-band filter as the V6 plate load. For a frequency in the 225.0- to 299.9-mc range, S1A selects the low-band filter.

The 2nd mixer, V7, functions as a band-pass amplifier or a mixer stage, as stated earlier. With the action of S1A to choose the high-band filter (as shown), another section of the switch allows the positive potential at the C37-R19 junction to be applied to the plate of V5.

It should be recalled that the V2 oscillator output (fig. 2-15) is also applied to the grid of the quintupler stage, V1. The plate tank of V1 (comprising L1 and C2) produces a flywheel effect at the fifth harmonic of the oscillator frequency (4.166667 mc). When the B voltage is applied to the V5 plate via S1A, the output is again multiplied in frequency doubler V5. The secondary of the V5 output transformer, T2, is tuned to the tenth harmonic of the 1st oscillator frequency, or 8.33333 mc. Thus, when the equipment is operating in the range from 300.0 to 399.9 mc, V7 receives the V6 output (20.3 to 21.3 mc) and the V5 frequency doubler output (8.33333 mc) at the control grid.

The tuned primary and secondary tanks of T3 are shunted by resistors to lower the Q of the circuits. The 12.0- to 12.9-mc output of the 2nd mixer stage is applied to the grid of the 3rd mixer, V9.

The 2nd crystal oscillator, V8, is connected as a crystal-controlled Colpitts circuit. The oscillator generates either of five frequencies, as shown in the block diagram (fig. 2-4). The plate of the oscillator, V8 (fig. 2-15) is held at r-f ground potential by C78. Feedback is developed between the grid and cathode by C77. Grid leak bias is derived from the charge-discharge action of the crystal holder capacitance of the chosen crystal through series resistors R25 and R26.

Rotation of the third digit clutch plate, CP2, by the motor during the monitor tuning process, drives switch section S2C to the position dictated from the memory cylinder. In this manner, either of the crystals (Y2 through Y6) is shunted across the grid circuit of V8A. Thus, V8A applies the desired frequency to the suppressor grid of V9, while V7 supplies the 12.0- to 12.9-mc input to the control grid.

The 8.8- to 9.0-mc output of V9 is fed through low Q band-pass filter circuits to the 4th mixer, V10, control grid. The 3rd crystal oscillator, V8B, generates either of ten frequencies (see fig. 2-4). The crystal required to produce the correct frequency (as determined by the memory cylinder) is selected as a result of the rotation of clutch plates CP2 and CP3 (fig. 2-15). CP2 drives switch section S2B, while S3C and S3D are being driven by CP3. The chosen crystal (Y7 through Y16) is connected between grid and ground of V8B through the contacts of the associated switch sections.

The operation of the 3rd crystal oscillator, V8B, is identical to that of the 2nd crystal oscillator just discussed. The 3rd crystal-oscillator output is applied from the V8B cathode to the suppressor grid of the 4th mixer, V10. The oscillator signal is combined with the 8.8- to 9.0-mc, band-pass filter output at the 4th mixer to produce a beat frequency of either 5.172917 or 5.181250 mc.

DISCRIMINATOR

The tuned primary and secondary of T5 constitute a band-pass filter, which passes the 5.1- to 5.2-mc output of V1. The signal is applied to the discriminator stage, V12. The discriminator also receives either a 5.172917- or 5.181250-mc input via the secondary of T6. The purpose of the discriminator is to indicate the phase deviation between the two inputs, and to convert any error in phase to a voltage that is proportional to the error. The error voltage

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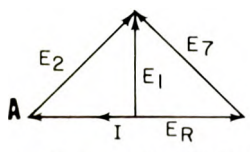
is applied to either the reactance tube, V5 (fig. 2-10) or to the motor control amplifier, V13 (fig. 2-15).

The 4th crystal oscillator, V11A, generates a reference frequency, which is applied to the discriminator, V12. The oscillator output frequency is 5.171917 mc for frequencies having an even fourth digit (as determined from the memory cylinder) and 5.181250 mc for frequencies whose fourth digit is an odd number. Switch section S3B driven by CP3 during motor rotation selects the proper oscillator crystal in accordance with the fourth digit of the frequency set on the memory cylinder.

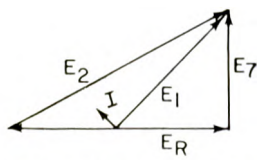
The chosen crystal (Y17 or Y18) selected by S3B is connected from grid to ground of V11A. The circuit operation is similar to that of the 2nd and 3rd crystal oscillators just discussed. The oscillator output is coupled from the cathode of V11A to the V11B grid by C92. V11B functions as a cathode follower (isolation amplifier), allowing its output to be developed across a tank circuit at the cathode. Cathode bias is obtained across R41 and C85. The oscillator output is coupled from the cathode tank to the discriminator input via the T-6 secondary.

The relative phase of the two input signals at the discriminator (from the 4th mixer, V10, and 4th crystal oscillator, V11) causes the circuit to produce an output voltage proportional to phase error. To understand this action, consider the simplified circuit diagram of figure 2-16. E1 represents the input voltage from the 4th mixer across L19, while E_r is the reference voltage from the 4th crystal oscillator at L21.

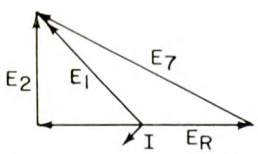
For zero error the 4th mixer output, E1, is approximately 90° out of phase with the 4th crystal-oscillator voltage, E_r. This relationship exists as a result of the L19 input being applied to the center tap of L21. L21 receives a voltage induced from the T5 primary. The resultant voltages, E2 and E7, applied to the plate of V12A and V12B, respectively, are equal in amplitude, as shown on the vector diagram (fig. 2-16, A). Note that V12A conducts a current up through its load component, R44, to develop a positive voltage at pin 5 with respect to ground. V12B conducts a current down through R45 to develop



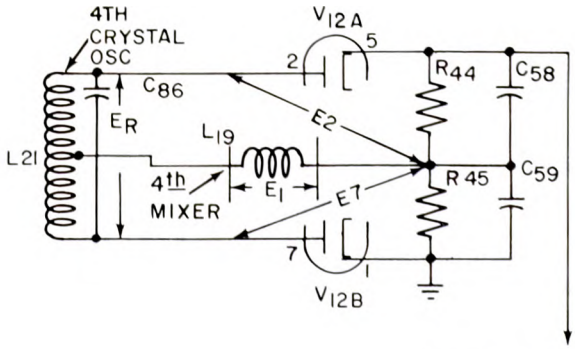
VFO ON FREQUENCY



B VFO TENDING TO GO LOW IN FREQUENCY



C VFO TENDING TO GO HIGH IN FREQUENCY



DISCRIMINATOR OUTPUT TO REACTANCE TUBE AND MOTOR CONTROL AMPLIFIER V13A (FIG. 2-18)

Figure 2-16. –Discriminator circuit.

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a negative voltage at the R45-R44 junction with respect to ground. The two voltages add algebraically across C58 and C59, respectively, at the output to produce a zero discriminator error voltage of +5 volts.

Vector displacement to produce the +5-volt, zero error output is not shown in figure 2-16. The reason for the selection of +5 volts to represent zero error is described later. The output of the discriminator is applied to a motor control amplifier (discussed later) and to the reactance tube in the r-f head (fig. 2-10).

If the 4th mixer output, E1, tends to lag its normal phase as a result of the vfo drift toward a lower frequency (fig. 2-16, B), voltage E2 becomes larger than E7, and the discriminator output is more positive to ground. Conversely, if the 4th mixer output, E1, tends to lead its normal phase due to a drift in the vfo output toward a higher frequency, the magnitude of voltage E7 becomes greater than E2 (fig. 2-16, C), and the discriminator output may go negative with respect to ground.

TUNING MOTOR CONTROL

A block diagram of the tuning motor control circuits is shown in figure 2-17. The vfo output sample is fed to the monitor. The monitor supplies an output to the clutch control amplifier and discriminator circuits. The discriminator output controls the magnitude of the conduction through a duo-triode motor control amplifier. The low and high discriminator relays determine the direction of rotation of the tuning motor to tune the vfo tuning capacitor.

The tuning motor controls the tuning capacitor of the vfo by three clutches designated the high-,

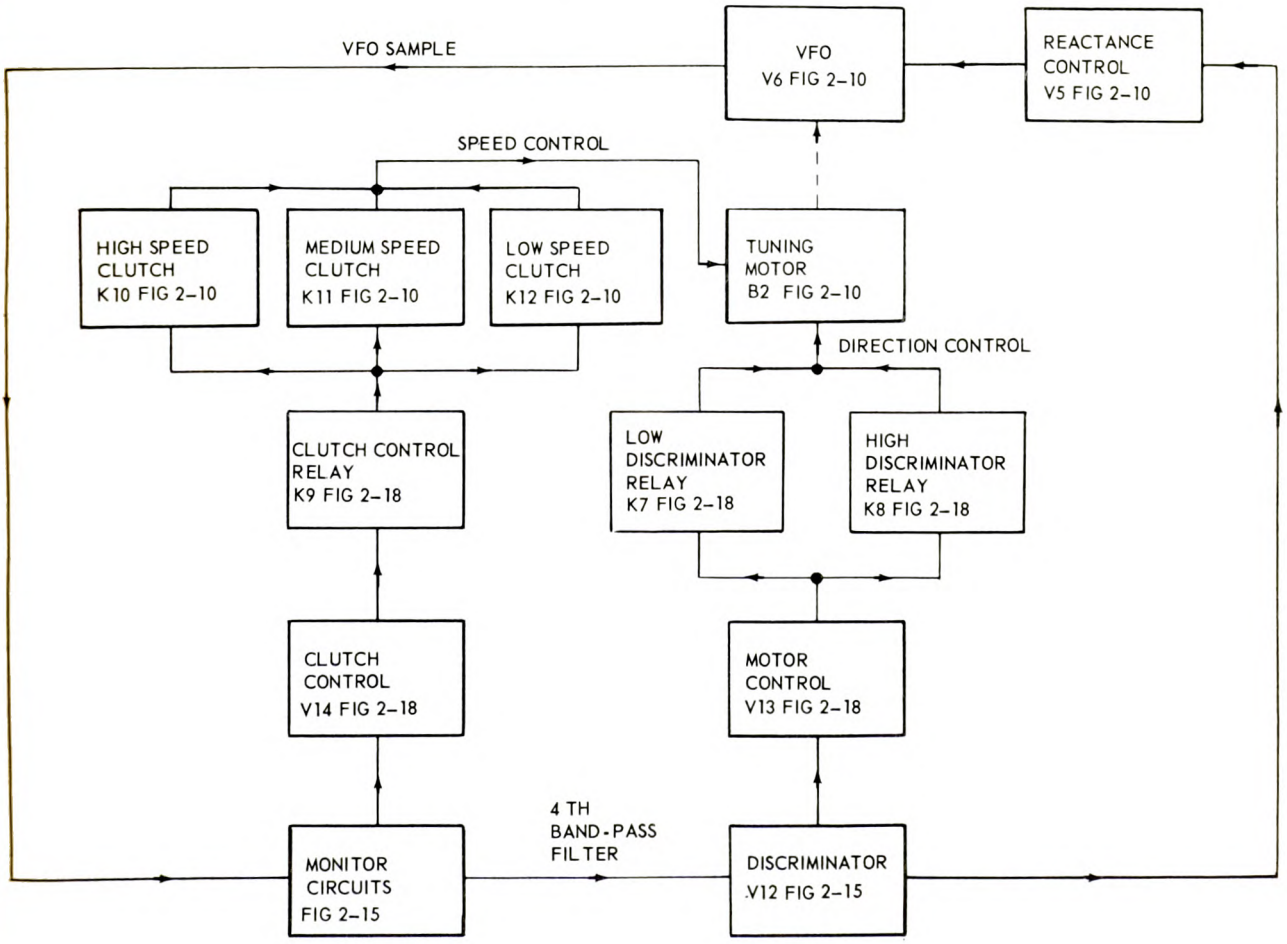

Figure 2-17. —Channel selection circuit.

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medium-, and low-speed clutches, respectively. The total frequency range of the vfo is divided into six equal ranges. The high-speed clutch drives the tuning mechanism at high speed until the segment is reached in which the selected frequency lies. The medium-speed clutch then takes control to drive the tuning capacitor at medium speed until the approximate range of the half power points on the response curve of the 4th mixer is reached. At this point, the low-speed clutch takes over and drives the tuning capacitor to the desired channel frequency. The various clutches are energized by a clutch control relay, which, in turn, is activated from a clutch control amplifier.

After the bridge circuits in the monitor are balanced (as discussed earlier) by the selector motor, B1 (fig. 2-15) to the desired frequency, the discriminator, V12 (fig. 2-17) produces either a positive or negative output voltage, which is fed to a motor control amplifier, V13. The discriminator signal is received at the V13A grid through a limiting resistor, R54 (fig. 2-18). The plate loads of V13A and V13B are the solenoids of K8 and K7, respectively. These relays are connected through R55 to a +200-volt supply. K7 is in the low discriminator output path, and K8 is connected in the high discriminator output circuit. Both grids are tied through voltage dividers to the B supply so that they are positive to ground, and both sections of V13 are conducting.

When the vfo is off channel, and the selector motor, B1 (fig. 2-15) is setting up the system for selection of a new frequency, the bridge relays (fig. 2-15) of the monitor are energized, and the cathodes of V13 are tied to +30 volts. This action biases both sections of V13 (fig. 2-18) beyond cutoff throughout the frequency selection period.

The amount of the plate current passed by each section of V13 (when the tube sections are conducting) depends upon the magnitude of the discriminator output. For zero error in frequency, the discriminator output is approximately +5 volts. This value of voltage was chosen to represent zero error since zero voltage for zero error can exist either when the input frequency is correct, or when the frequency at the monitor input is too far away from the desired frequency to be passed by the monitor circuits.

Both tube sections of V13 conduct through common cathode resistor, R53. When the monitor circuits are on frequency, the discriminator output is +5 volts. The grid (pin 7) voltage is +8 to ground, while the cathode is approximately 10.2 volts. The V13B grid (pin 3) is about 5.3 volts to ground. Thus, V13B receives about -5 volts bias and does not conduct a sufficient current to energize its plate relay, K7.

When the input frequency is too far above the chosen frequency to be passed through the monitor (as may be the case when a new channel is selected), the discriminator output voltage is zero. The V13A grid voltage decrease from +5 volts to zero causes a smaller current through K8 (in the V13A plate circuit), and the K8 contacts are released (position shown). Likewise, the smaller current through R53 reduces the V13B bias. The increased V13B current through the solenoid of K7 energizes this relay to move the contacts down (opposite to that shown). This action causes the tuning motor, B2, to turn by completing the motor circuit from ground at the 2-4 contacts of K7 (K7 energized), through S4A (in the position shown), through the motor filter, and through the motor from a to b, through another motor filter, and the contacts of S4B (in the position shown), through the 8-7 contacts of K8 (K8 deenergized), through R56, the 3-2 terminals of K14, through the 3-4 contacts of K13, and to the +30-volt unfiltered motor power. (The motor power circuit is treated in detail later). The polarity of the voltage applied to the motor is such that the vfo tuning capacitor (C36 of figure 2-10) is driven towards the selection of a lower frequency.

When the 4th mixer, V10, output (fig. 2-15) is near the chosen frequency, the discriminator, V12, circuit is functioning, and the reactance tube (V5 of figure 2-10) begins to operate. The discriminator, V12, output (fig. 2-15) rises to approximately five volts positive. The conduction of V13A (fig. 2-18) increases, and its plate relay, K8, becomes energized. V13B continues to conduct, and both relays are now energized. The 8-7 contacts of K8 are opened when K8 is energized, while the 4-2 contacts of K7 (when the relay is energized) are closed. With S4A in the position shown this action removes drive power from the motor.

If the frequency at the discriminator (V12 of figure 2-15) input tends to fall below the zero error frequency (vfo low in frequency) the magnitude of the discriminator, V12, output voltage will be increased in the positive direction. V13A conducts heavily, and K8 is energized.

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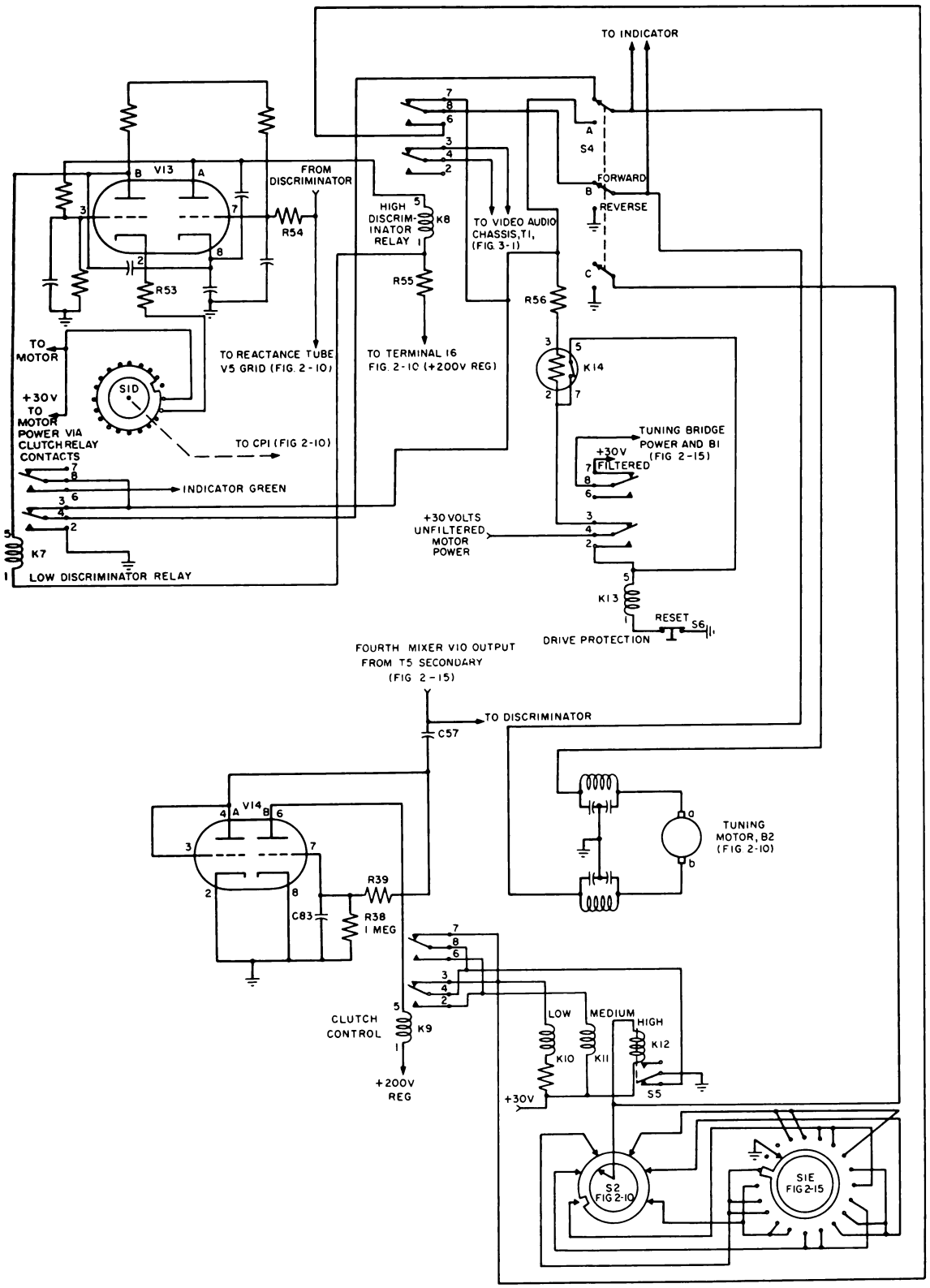

Figure 2-18. –Tuning motor control circuits.

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The bias (negative grid-cathode voltage) increase on V13B (as a result of increased current through R53 conducted by V13A) causes the V13B plate current to decrease, and K7 is deenergized. Power is now applied to the motor from a ground on the contacts of a high speed clutch control relay, K12 (discussed later). The motor circuit is connected from a ground on S5, through the 7-8 contacts of K9, through the 6-8 contacts of K8 (closed in this error condition) through the contacts of S4B (in the position shown), and through the motor filter circuit to "B" on the motor, B2. The +30-volts power is applied to the motor through the upper motor filter, through the S4A contacts (in the position shown) through the 3-4 contacts of K7 (deenergized for this error condition) through R56, K14, and the 3-4 contacts of K13, to the motor power. The direction of the current through the tuning motor, B2, is reversed for this error condition. Thus, the motor will run in the opposite direction until the discriminator output is again approximately +5.

From this action, it should be seen that when K7 is energized, a d-c voltage is applied to the tuning motor of such polarity that the vfo is driven (by its tuning capacitor C36 of figure 2-10) towards a lower frequency. If relay K8 (fig. 2-18) is energized, the vfo is driven towards a higher frequency. Both relays are energized when the vfo frequency is correct, and the tuning motor is braked at this point.

Forward-reverse switch S4 is cam operated, and effective in controlling the speed of the tuning motor. When the low end of the tuning capacitor is reached, the contacts of S4 are moved downward, and section C of S4 grounds one end of the high-speed clutch (fig. 2-18), thus energizing K12. When the high-speed clutch is energized, the tuning motor (and consequently the tuning capacitor C36) is driven at a high speed toward the high frequency end of the tuning range. Note that the conduction path for B2 is from a ground on S4B and through the motor from b to a when S4 is in the down position.

When the tuning capacitor reaches the high end, the cam (not shown) that operates S4 has rotated so that S4 is released, and motor power is now applied so that the motor current is from a to b, thus causing the motor to turn in the reverse direction. The motor continues to turn at high speed since K12 remains energized by a ground return supplied by segment selector switch section S1E.

The action of S1E is to divide the entire frequency range into six frequency segments. When the high-speed clutch, K12, is energized, the tuning motor continues to operate at high speed until the sector is reached in which the preset frequency lies. At this time S1E (which is driven by CP1 of figure 2-15) operates to remove the ground from the high-speed clutch, K12 (fig. 2-18). When the tuning system is far off frequency, K9 is energized with its 4-2 and 8-6 contacts closed. This action will be considered later. The deenergizing of K12 permits S5 to supply a ground for the medium-speed clutch, K11. The K11 circuit is completed from ground, through the lower contacts of S5, through the 4-2 and 8-6 contacts of K9 in parallel through the K11 solenoid, to the +30-volt motor power. With K11 energized, the tuning motor is driven at medium speed until the selected frequency begins to pass through the monitor.

When the output of the 4th mixer (V10, figure 2-15) is such that it can pass through the fourth band-pass filter circuit comprising the tuned T5 primary and secondary, a signal is coupled by C57 to the clutch control amplifier, V14. V14A is connected as a diode (fig. 2-18), which rectifies the V10 output. Each time the input signal goes positive with respect to ground, V14A conducts to effectively short R39 and R38. Negative portions of the signal are developed across R38 and R39, and applied from the junction of the two resistors to the control grid of V14B.

The negative pulses at the grid of V14B cause a decrease in the V14B plate current. The smaller current through the K9 solenoid deenergizes this relay, and the contacts are released (as shown). With K9 released, the ground is removed from K11 and supplied to the low-speed clutch, K10, through the 3-4 contacts of K9. This action causes the motor to turn at low speed until the exact frequency is reached.

When searching for the desired frequency, the tuning capacitor, C36, travel (fig. 2-10) is always from the high to the low frequency end of the tuning range. If a new channel, which lies between the previous channel and the low frequency end of the tuning range, is selected, the tuning motor will be driven to the new channel and stopped. The tuning drive may start at high, medium, or low speed, depending upon how far removed the new channel lies from the old.

If the channel is far removed, S2 and S1E (fig. 2-19) ground K12, and the motor turns at

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high speed. If K12 remains deenergized when the new frequency is selected, S5 grounds K11 (through the 2-4 contacts of K9), and the motor turns at medium speed. If the new frequency is such that the monitor continues to receive an input, K9 will be deenergized as a result of negative pulses at the V14B grid (which reduces the V14B plate current), and the ground from S5 is supplied through contacts 3-4 of K9 to the low-speed clutch. Thus, the motor will turn at low speed to the new frequency.

If a new channel frequency, which lies between the previous frequency and the high frequency end of the tuning range is selected, the tuning capacitor, C36, will first be driven to the high frequency end. Cam operated switch S4 then reverses the direction of rotation of the motor, causing the motor to start back again to search for the new channel frequency.

In the event that the vfo (fig. 2-10) is not set to the new channel frequency after traveling completely across the tuning range, it is returned to its high frequency end at high speed by the reversal of S4 (fig. 2-18). In the reverse position, S4A and S4B connect the motor directly in series with R56, K14, and the 3-4 contacts of K13, across the +30-volt power by passing relays K7 and K8 and thus remove the motor entirely from the control of the monitor.

When the vfo reaches the high frequency end of its range, S4 returns to the position shown. The monitor circuit discriminator relays, K7 and K8, again take control, and the tuning capacitor, C36, is driven back towards its low frequency end in search of the new channel frequency.

A thermal time delay relay, K14, protects the tuning motor, B2 (fig. 2-10) the drive mechanism, and the selector motor, B1 (fig. 2-15) in case of repeated futile attempts to select the new frequency. Drive protective relay K13 is normally deenergized as a result of the open 5-7 contacts of K14. After several trips across the tuning band, heat produced by the resistance between the 2-3 terminals of K14 causes the 5-7 contacts of K14 to close, and power is applied to K13. Thus, K13 energizes, and motor power for B1 and B2 is removed by the opening of the 4-3 contacts of K13.

To locally place the motor in operation again at the receiver, a reset switch, S6, must be depressed. This action opens the power circuit to the K13 solenoid, and the relay contacts release to again apply power to the motor. A similar reset switch is situated on the front panel of the indicator (fig. 2-1), which permits the motor power to be restored from a remote location. This switch performs the same function as that of S6.

QUIZ

1. (a) What is a direction finder? (b) What information is presented by a direction finder?
2. What are the three major units of the direction finder, AN/URD-4?
3. The antenna of the AN/URD-4 direction finding system is of what type?
4. Direction finding information is presented in what form on the face of the cathode-ray tube of the indicator?
5. R-f antenna voltage is fed to the receiver through a balance-to-unbalance device known as what type circuit?
6. (a) The purpose of the monitor (fig. 2-2) is to control the frequency of what unit? (b) The monitor achieves its purpose by controlling what components?
7. The harmonic generator (fig. 2-4) produces frequencies ranging between what harmonics of the oscillator fundamental frequency?
8. If the monitor (fig. 2-4) is set to a frequency in the 225.0- to 299.9-mc range, a selector switch permits the output of the 1st mixer to be fed through what type filter?
9. The 10th harmonic of the 1st crystal oscillator frequency (fig. 2-4) is produced by the action of what two circuits?
10. A 2nd crystal oscillator (fig. 2-4) provides a choice of how many frequencies?
11. How many frequencies are fed to the discriminator circuit (fig. 2-4)?
12. The discriminator (fig. 2-4) is a phase sensitive device, which develops an output voltage representing the resultant of what deviation between the two frequencies?
13. The discriminator output (fig. 2-4) is fed to what circuit?
14. During channel selection (fig. 2-5) the setting of what component determines which 20 of 1750 available frequencies will be selected by remote control as the (preset) channel frequencies?
15. The rotating antenna array (fig. 2-8) is coupled to the stationary portion of the antenna through what components?
16. The antenna drive motor, B1 (fig. 2-8) is an induction-type, single-phase motor, which drives a single antenna shaft at what speed?

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17. The signal from the antenna is fed via P1 (fig. 2-8) to what portion of the receiver (fig. 2-10)?
18. The center post of the cavities (fig. 2-11) acts as what type component?
19. The variable frequency oscillator circuit V6A (fig. 2-12) is connected as what type oscillator?
20. Z1 (fig. 2-10) comprises high Q, double-tuned, parallel tank circuits whose peak response occurs at what frequency?
21. When one of the preset frequencies is to be selected, the channel selector switch is turned to the new position which causes what acticn?
22. A change in the input frequency to the r-f head (fig. 2-10) necessitates a change in what other circuit frequencies?
23. The channel selecting bridge circuits (fig. 2-13) are established between what two units?
24. What capacitors (fig. 2-15) make possible the generation of any harmonic of the fundamental frequency from the 4th to the 13th harmonic?
25. The heterodyning processes in the monitor (fig. 2-15) should always produce either of what two frequencies?
26. The tuned primary and secondary tanks of T3 (fig. 2-15) are shunted by resistors for what reason?
27. The crystal required to produce the correct frequency for the 3rd crystal oscillator (fig. 2-15), as determined by the memory cylinder, is selected as a result of the rotation of which clutch plates?
28. The relative phase of the two input signals at the discriminator (fig. 2-15) causes the circuit to produce what type output voltage?
29. The monitor (fig. 2-17) supplies an output to what two circuits?
30. The low and high discriminator relays (fig. 2-17) have what effect on the motor that tunes the vfo tuning capacitor?
31. The tuning motor (fig. 2-18) is caused to control the vfo tuning capacitor by means of what components?
32. After the bridge circuits in the monitor are balanced by the selector motor, B1 (fig. 2-15) to the desired frequency, the discriminator produces either a positive or negative output voltage, which is fed to what circuit?
33. The amount of the plate current passed by each section of V13 (fig. 2-18) depends upon the magnitude of what circuit output?
34. For zero error in frequency, the discriminator output (fig. 2-15) is approximately what value?
35. When the input frequency (fig. 2-15) is too far above the chosen frequency to be passed through the monitor, the discriminator output voltage is what value?
36. When the 4th mixer output (fig. 2-15) is near the chosen frequency and the discriminator is functioning, what circuit begins to operate?
37. When K7 (fig. 2-18) is energized, a d-c voltage is applied to the tuning motor of such polarity that the vfo is driven towards what frequency?
38. When searching for the desired frequency, the tuning capacitor travel is always in what direction?
39. A thermal time delay relay (K14 of figure 2-18) protects what components in case of repeated, futile attempts to select the new frequency?

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CHAPTER 3

ELECTRONIC AIDS TO NAVIGATION: RDF (PART II)

INTRODUCTION

The radio direction finder is a highly sensitive receiver to which a directional antenna is connected. Direction finding information is presented on the face of a cathode-ray tube in the indicator as a propeller-shaped pattern. The pattern may be transformed to a distorted figure 8 shape with its apex pointing to the direction of the received signal.

The circuit operation of the i-f video-audio and the indicator chassis of the Radio Direction Finder, AN/URD-4, are considered in this chapter. A block diagram treatment of these units, along with the circuit operation of the antenna, r-f head, and monitor, was given in chapter 2 of this training course. It may be necessary on occasion to refer to the material of chapter 2 in order to better understand the contents of this chapter.

CIRCUIT OPERATION

I-F VIDEO-AUDIO CHASSIS

The primary purpose of the i-f video-audio chassis located in the receiver cabinet (chapter 2, figure 1) is to amplify and detect the i-f signal. The chassis also provides automatic gain control (AGC) circuits for application of control voltage to the r-f and i-f stages. A squelch circuit disables the audio stages in the absence of a received signal.

A block diagram of the complete receiver is shown in chapter 2, figure 4. It should be recalled from the block diagram discussion that the 15.325-megacycle i-f signal from the i-f amplifier, V9, in the r-f head is applied to the second mixer stage, V1. The 15.325-mc input is heterodyned with a 12.5178-mc signal from the crystal oscillator, V2. The mixer output frequency (2.8072 mc) is amplified in three stages of the i-f amplification. A portion of the third i-f input is applied to an AGC amplifier, V7, and subsequently rectified to form a d-c AGC control voltage. AGC control is active at the grid of all the r-f and i-f stages.

The third i-f output is rectified and fed along dual paths to the audio and video circuits, respectively. The audio signal is amplified in three stages of audio amplification (two voltage and one power amplifier). The audio output is made available at the headphone jack on the front panel of the receiver chassis (chapter 2, figure 1).

The video output from the signal rectifier (detector) is fed via a cathode follower (for impedance matching purposes) to the indicator. A high-pass filter circuit (chapter 2, figure 4) couples noise impulses to a noise amplifier. These signals are amplified, and subsequently rectified, for application to the indicator as noise compensation voltage.

Input Stages

The 15.325-mc signal at J3 (fig. 3-1) arriving from V9 in the r-f head of the receiver (chapter 2, figure 10) is coupled from Z3 to a second mixer stage, V25. This signal is applied to the control grid of the stage.

A Pierce oscillator, V26, which is connected as a triode to increase interelectrode capacitance for feedback, generates a 12.5178-mc signal. The oscillator output is coupled by C88 to the suppressor grid of V25, and is heterodyned with the 15.325-mc input from the r-f head. The resultant difference frequency (2.8072 mc) is developed across the primary tank of Z4.

I-F Stages

The difference frequency is amplified in three conventional stages of i-f amplification, comprising tubes V27, V28, and V29. These stages are cascade transformer coupled, having their peak response at approximately 2.8 mc.

The input to the third i-f amplifier stage, V29, is applied simultaneously to V29 (through C122) and to an AGC amplifier, V31, through

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R211. The input signal is amplified by V31 and developed across the plate load, L11. C147 then couples the V31 output via a tuned circuit, L12 and C148, to the plate of the AGC rectifier, V32B. The Q of the V32 input tank is decreased by the loading presented by V32B.

Automatic Gain Control

The AGC Circuit functions as follows: A positive bias of 21 volts is applied to the V32B cathode from the R217-R218 junction of a voltage divider comprising R217, R219, and L18 connected between the +200-volt, regulated supply and ground. Signals applied at the V32B plate must exceed the +21-volt cathode (pin 1) potential in the positive direction to produce a V32B, AGC output current.

When the input signal is of sufficient amplitude, the AGC rectified current from the V32B plate is conducted through L12, R215, R214, R213, and through the r-f gain control potentiometer, R250, to ground. The series circuit comprising R213 and R250 is in parallel with R220 to ground. C149, C150, C151, and C152 filter the AGC output so that the AGC voltage is effectively a constant negative potential to ground. Additional filtering is provided for the AGC voltage applied to the V29 grid by R210 and C144.

A less negative AGC potential than that applied to the V29 grid is applied from the R213-R214 junction (through the filter components C152, R212, and C151) to the grids of V25, V27, and V28. This potential is also applied to an i-f amplifier stage, V9, in the r-f head (see figure 2-10, chapter 2).

A separate AGC rectifier, V32A (fig. 3-1), produces the automatic gain control voltage for the r-f stages in the r-f head. This section of V32 receives a larger value of fixed voltage at its cathode (approximately +30 volts from the R221-R218 junction to ground) and therefore requires a larger amplitude signal to cause the production of an AGC voltage for the r-f stages. In this manner, only the larger amplitude signals arriving at the r-f input (figure 10, chapter 2) are affected by the automatic gain control voltage.

The AGC amplifier (V31) output is received at the V32A plate via C156. The output is also applied through R222 to the r-f AGC clamp diode, V37. The plate of the clamp diode is tied through R221, R218, and L18 to the B supply. Thus, V37 normally conducts to hold the R221-C158 and R222-R223 junctions slightly positive to ground. The positive potential (about 0.5 volt) is applied from the filter network comprising R223 and C157 to the control grids of the r-f stages (chapter 2, figure 10) during the reception of weak signals.

For strong r-f input signals, the input to the AGC amplifier, V31 (fig. 3-1) will be of larger amplitude. Likewise, the output of V31 to the r-f, AGC rectifier, V32A, will be of sufficient amplitude (in the positive direction) to cause the V32A conduction. For positive portions of the input, the V32A conduction will cause C156 to charge negative toward the plate (pin 2). During negative portions of the input from V31, V32A will not conduct. C156 will discharge through R223, C157, and through C149 and R216. The time constant formed by the filter components, C157 and R223, is sufficiently long, relative to the period of the input frequency to maintain the C157 charge negative to ground for the duration of the reception of strong r-f signals. Thus, the negative voltage developed by C157 holds the r-f stages control grids negative during the application of strong signals to V32A. The action of both sections of V32 to produce the AGC voltage is that of "delayed AGC."

The effect of the AGC voltage to decrease the gain of the r-f stages can be controlled by adjusting the position of the arm of the r-f gain control, R250. The control is located on the panel of the indicator (chapter 2, figure 1).

A gain control cathode follower, V36 (fig. 3-1) is connected between a +200-volt supply (connected to the plate) and a -150-volt supply (connected to its cathode through R241). A portion of the negative rectified voltage from the i-f AGC diode, V32B, is fed through the local-remote switch (section S2B in the position shown) and developed across R250. This potential is applied to the V36 grid. The amount of the negative potential is determined by the setting of the arm of R250.

If the arm of R250 is moved upward, the negative voltage to the grid of V36 is decreased. V36 conducts a heavier current to develop a more positive voltage at the V36 cathode via R241. The change in the cathode potential is applied through R240 to the plate of the r-f AGC clamp diode, V37.

The V36 cathode potential may be either positive or negative (depending on the grid potential) since V36 is connected between the -150- and +200-volt supplies. If the cathode voltage is positive, V37 conducts to clamp the voltage a few volts positive to ground, and the

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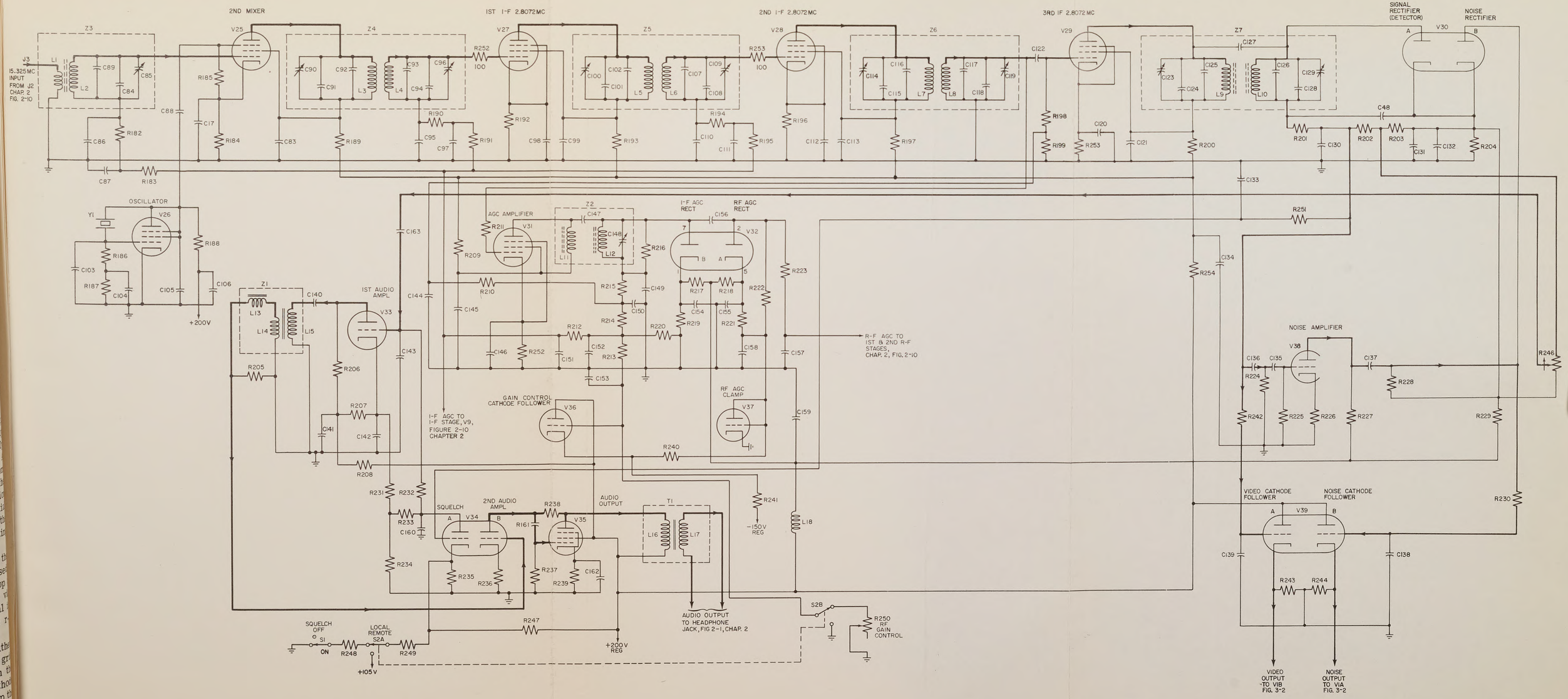

Figure 3-1. -I-f video audio chassis.

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gain of the r-f stages is increased. If the V36 cathode potential is reduced but remains negative to ground, V37 does not conduct, and the less negative cathode potential of V36 is applied through R240, R222, and R223 to the r-f stages. Thus, the gain of these stages is increased.

Conversely, if the arm of R250 is moved downward, the grid of V36 is made more negative and the conduction of V36 decreases. The V36 cathode potential is made less positive. If the potential still remains in the positive region, V37 continues to clamp the r-f AGC voltage a few volts positive, and the r-f stages are not greatly affected. If the voltage becomes more negative to ground, the greater negative cathode potential of V36 is applied to the r-f stages, and the gain of these stages is reduced.

Signal Rectifier

The signal rectifier (detector, V30A in figure 3-1) receives the 2.8072-mc input from the third i-f stage, V29, via the double-tuned transformer of Z7 and a pigtail coupling capacitor, C127. The bandpass of the Z7 tank is approximately 50 kc. Since the capacitive reactance of C127 decreases with a frequency increase, C127 compensates for losses in the transformer coupling at high frequencies, thereby increasing the amount of coupling for the higher frequency components of the i-f signal.

The plate and cathode of V30A complete a circuit through the input tank, which supplies the source voltage, and through R201, R202, and R203. Positive alternations of the input cause a current through these components whose amplitude is determined by the positive value of the separate alternations. Likewise, the load voltage developed across R201, R202, and R203 is also a function of the input signal amplitude. The detector output video contains the rectified audio signal and noise frequencies. The audio voltage is used for aural monitoring. The noise frequencies developed across the detector load components are subsequently rectified and used to establish a d-c bias level, which determines the amplification of the video signal. This action is discussed later.

The entire V30A output is held above ground potential by the amount of the voltage across R204. R204 connects in series with R229 and L18 to the B supply.

That portion of the audio output developed across R203 is also available across an audio level control, R246. The R246 arm may be varied to control the amount of the total audio output voltage, which will be applied through C163 to the first audio amplifier (V33) control grid.

First Audio Amplifier and Squelch Circuits

The first audio amplifier, V33, has its grid returned to ground through R232, R233, and R234. The output from the signal rectifier, V30A (through C163) is developed at the V33 grid across these resistors, and filtered by C143.

A positive voltage is applied from the R207-R231 junction to the cathode of V33. These resistors form a portion of a voltage divider (comprising R208, R207, R231, and R234), which is connected from the +200-volt B supply to ground. R233 is connected from the R231-R234 junction to the plate of the squelch tube, V34A. The plate is returned through R233, R231, R207, and R208 to the +200-volt supply.

In the absence of an input signal, the voltage at the V34A cathode is approximately equal to the voltage at the grid. The potentials are obtained in the following manner: a squelch switch on the indicator panel (chapter 2, figure 1) connects two resistors, R249 and R248, from ground to the V34A cathode (fig. 3-1) when the switch, S1, is in the ON position, as shown. This action decreases the cathode-to-ground resistance below that of R235, and a smaller grid-cathode bias voltage for V34A is produced. In this condition, the cathode voltage is approximately 16 volts. Likewise, the value of voltage developed across R204 (applied to the grid of V34A through R203, R202, and R251) is also about 16 volts. Thus, the bias voltage of V34A (with the switches at the cathode as shown) is almost zero, and the tube is conducting when no signal is received.

The V34A plate current through R233 increases the positive potential at the cathode of V33, and increases the negative (bias) voltage at the V33 grid. This action makes the V33 grid about -38 volts to the cathode, which biases V33 beyond cutoff, and the audio circuits are muted.

When a carrier signal is detected by the signal rectifier, V30A, a negative (or less positive) potential is fed to the grid of squelch tube V34A, causing the V34A current to decrease to cutoff. The current through R233 drops to zero, and the grid-cathode bias of V33 is reduced to that value of voltage existing across R231. This potential allows V33 to conduct and to pass the audio signal supplied to its grid through C163.

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When the squelch switch, S1, is placed in the OFF position, the cathode resistance of V34A rises, and the cathode voltage rises to approximately +28 volts. The grid potential remains at +16 volts, and the grid is therefore -11 volts with respect to the cathode. Thus, V34A is cut off when S1 is open, and the grid-cathode voltage of V33 (developed across R231) operates the tube along the linear portion of its eg-ip curve to amplify the audio signal. More negative signals from V30A drive the grid of V34A further into cutoff, and the squelch control is entirely removed from the audio circuits.

A local-remote switch is situated behind the front panel cover of the receiver chassis. (See chapter 2, figure 1.) When the switch, S2A (fig. 3-1) is placed in the LOCAL position, +105 volts is directly applied to the V34A cathode, which also removes the control of the squelch circuit.

Z1, in the plate circuit of V33 (fig. 3-1) allows maximum passage of frequencies between 150 and 4500 cps. The output inductor, L14, of Z1 is frequency compensated by L13 and R205. L13 is a series placing coil that resonates with the output capacitance and increases the upper frequency limit. R205 lowers the circuit Q and also serves as the grid resistor for the V34B stage which follows.

Second Audio Amplifier

The second audio amplifier, V34B, receives the audio output of V33. A grid-cathode bias voltage (1.4 volts) is developed by V34B as a result of the cathode current through R236. This voltage establishes the operating point of V34B in the linear portion of its characteristic curve to allow faithful reproduction of the input signal. R236 is unbypassed, and is therefore degenerative towards the grid input to further increase fidelity. Due to the large V34B plate load and the resulting small Ep, the plate current is small, and V34B functions as a voltage amplifier.

Audio Output Stage

The audio output stage, V35, is also operated class A. Its cathode voltage across R239 is not degenerated to allow a larger V34B output signal. The output signal is further increased as a result of the relatively high screen grid potential (slightly higher than the plate) to increase the stage gain.

The audio output of V35 is coupled by T1 to the headphone jack on the receiver front panel. When the tuning system (discussed in chapter 2) is tuning the r-f head toward a lower frequency relay, K8 (chapter 2, figure 18) is deenergized. Thus, the 3-4 contacts of K8 short the T1 secondary to mute the audio (V35) output (fig. 3-1). Although the squelch circuit may be active to produce an output during the time of low frequency tuning, no signal will be presented to the headphones. The time of the tuning from the low to the high frequency end of the tuning range is relatively short, requiring only a few seconds. The video output is not affected by the action of the squelch circuit.

The video component of the detected signal from V30A at the R202-R251 junction (fig. 3-1) is applied to two paralleled stages, which are a noise amplifier, V38, and a video cathode follower, V39A. The cathode follower (V39A) impedance matches the output of V30A to the video input stage of the indicator (discussed later). The noise amplifier, V38, supplies a voltage to the video differential amplifier, V1A (fig. 3-2), the amplitude of which is proportional to the d-c level of the noise input frequencies.

Noise frequencies in the detected signal are generally of a higher frequency than the video component. Note that the video input voltage to the V39A grid (fig. 3-1) for each component of the input signal is that value of the separate voltages to which C139 is allowed to charge. For noise (high) frequencies, C139 will charge to only a very small amount of the voltage, since the duration of the separate noise pulses is small. On the other hand, the lower frequency video component of the V30A output exists at C139 for a longer period, and the capacitor develops a larger amount of the video input voltage (at the V39A grid) than that of the noise voltage. The cathode load resistors for V39A and B (R5 and R1, respectively) are located in the indicator chassis (fig. 3-2). Due to the action at the grids of V39A and V39B, as described in figure 3-1, the major portion of the V39A cathode output is the video signal, which is fed to the grid of V1B (fig. 3-2).

Noise Amplifier Circuits

The noise amplifier, V38 (fig. 3-1) functions in the following manner. All of the V30A voltage output is developed across the R242-C139 voltage divider. Only the higher frequency components are developed across R242 (due to the low impedance of C139 at these frequencies), and this voltage is applied through C136 and C135 to the noise amplifier (V38) grid. For lower (video)

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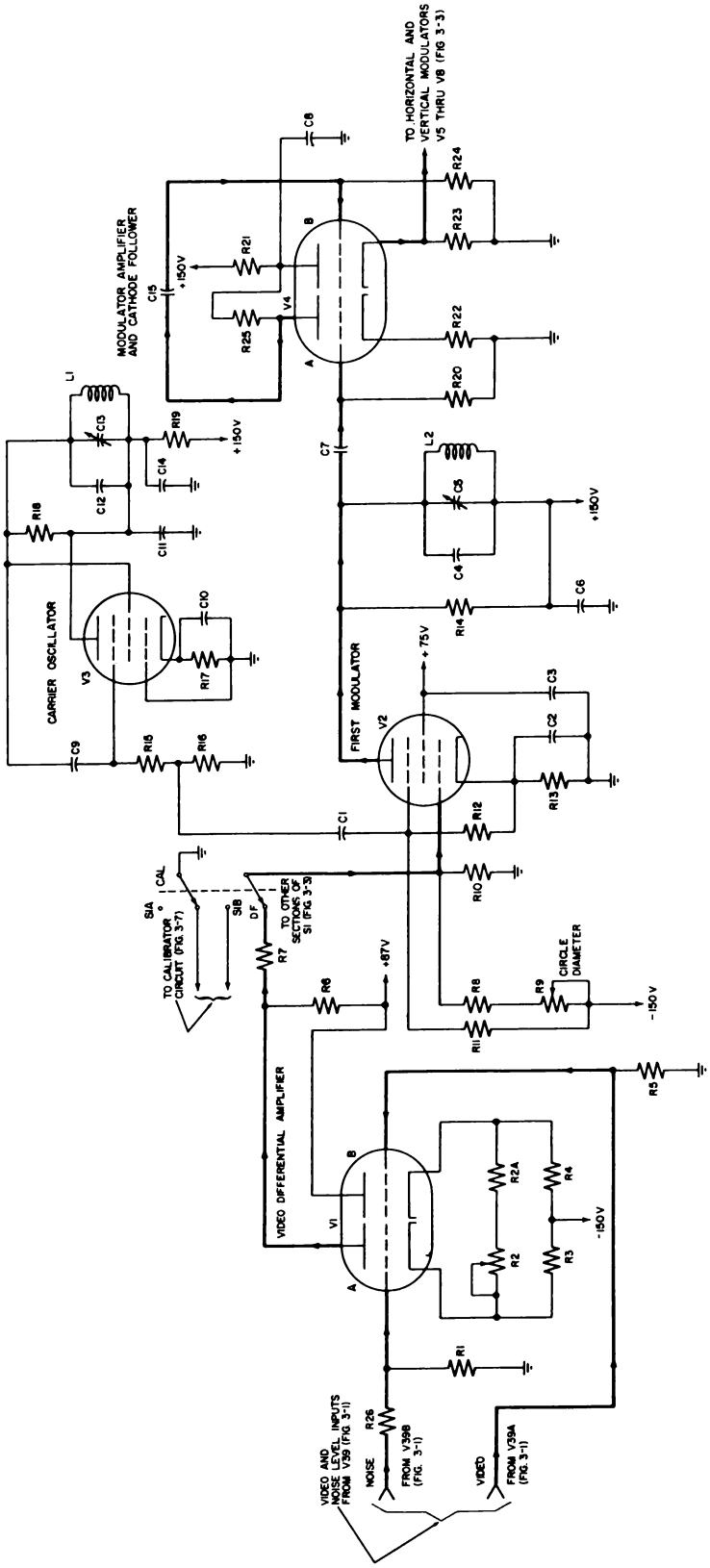

Figure 3-2. -Azimuth indicator (video amplifier and first modulator circuits).

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frequencies, the total V30A output is divided between C139 and R242, and the amount of the lower frequencies coupled by C136 and C135 to the noise amplifier grid is decreased. Thus, the greater portion of the input to V38 is noise, or higher than video frequencies. C136, R224, C135, and R225 filter the noise signal so that the V38 grid input is generally a voltage proportional to the magnitude of the noise signal input.

Noise rectifier V30B receives the V38 output via C137. V30B is normally conducting. Positive portions of the noise pulses cause a heavier V30B current, and a drop in the V30B plate potential. The voltage changes are impressed across R230 and C138 to ground. That value of the output at the R230-C138 junction serves as the input signal to the noise cathode follower (V39B) control grid.

The noise cathode follower is normally conducting as a result of a small positive potential across C138 applied to its grid. C138 forms a portion of a voltage divider (comprising C138, R230, R228, R229, and the filter choke, L18), which is connected from the +200-volt B supply to ground.

Strong signals from V38 cause V30B to conduct more current through R228, R229, and L18 to the B supply. This action lowers the positive voltage at the R230-R228 junction. This change in voltage is coupled through R230 to C138 and thus lowers the positive voltage on the V39B grid. V39B conducts less current through R244. The R244 cathode voltage is proportional to the magnitude of the noise voltage. It should be noted that there is essentially no a-c components in the voltage across R244.

INDICATOR CHASSIS

The indicator chassis of the AN/URD-4 (chapter 2, figure 1) provides the means by which the detected video signals (from the i-f video-audio chassis, figure 3-1) are amplified, mixed with the sweep voltages originating at the antenna, and presented to the cathode-ray tube for display. A block diagram consideration of the indicator stages and their functions was treated in chapter 2 of this training course. You should refer to that discussion for a brief review of the interaction of the indicator stages before attempting to understand the material presented here.

Video Differential Amplifier

The video differential amplifier, V1 (fig. 3-2) receives the video input signal from V39A (fig. 3-1) at its B section grid (fig. 3-2). At the same time, a d-c cathode potential proportional to the level of the noise is applied to the A section grid of V1 from V39B (fig. 3-1). Both sections of V1 are connected between the -150- and +87-volt supplies.

The bias applied to each section of V1 (fig. 3-2) is approximately 1 volt, and both sections of V1 are normally conducting. If the noise component of the signal to V39B (fig. 3-1) increases, the positive charge across C138 decreases (as discussed), and the cathode-to-ground potential of V39B (across R1, figure 3-2) decreases. This action lowers the conduction of V1A (fig. 3-2). Note that the cathode-to-ground potential will decrease for both sections of V1 since the two cathodes are tied together by R2 and R2A. R2 can be adjusted to balance the currents of V1A and V1B in the absence of an input signal.

The video signals to the V1B grid (from V39A, figure 3-1) are negative (or less positive going) pulses from V30A. For zero video input to V39A (or when the received video pulse amplitude reduces to zero) V39A conducts heavily, and the positive voltage at the V1B grid (across R5) is increased. V1B conducts heavily, and the cathode potential of both sections of V1 increases. Due to the increased common cathode potential, V1A can only conduct a small current through R6. The less positive potential at the top of R6 is applied through R7 and S1B (in the position shown) to the control grid of the first modulator, V2. Since the V1A conduction is small, the corresponding negative-going voltage to the V2 grid is not sufficient to cut off V2.

As the video signal increases in negative amplitude, V1B conducts less. The resulting decrease in cathode potential of both sections of V1 allows V1A to conduct a heavier current through R6. This action decreases the V1A plate voltage and the V2 grid voltage. The decrease (less positive) control grid potential at V2 cuts off the first modulator stage. Thus, for strong portions of the video (corresponding to maximum received signal at the antenna) the first modulator, V2, is held beyond cutoff to produce zero output. For zero video input, which corresponds to the nulls of the video, the first modulator stage, V2, is above cutoff,

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and thereby produces an output (as discussed later).

First Modulator and 200-KC Oscillator

The first modulator stage, V2 (fig. 3-2) provides a method of combining the video signal with an internally generated r-f signal to produce a modulated video output. The modulation process is necessary so that the proper mixing of the video and alternator voltages from the antenna can be accomplished in subsequent modulator stages.

The d-c potential at the grid of V2 is determined by the setting of the arm of circle diameter control, R9. The V2 grid potential, in turn, determines the operating level of the modulator, and likewise, the conducting or nonconducting condition of the modulator as discussed.

In the SENSE-D/F position of S1, the signal from the video differential amplifier, V1, is applied across R10 to the V2 control grid. When the video input signal at V1 is zero, V2 is conducting. A 200-kc carrier oscillator (V3 of figure 3-2) is connected as a transitron circuit. The transitron oscillator utilizes the negative transconductance of the tube to establish oscillations and develop an output frequency. Note that the screen potential of V3 is higher than that of the plate.

Screen grid signal current (which varies across C9, R15, and R16 at the frequency of the oscillator) causes a negative voltage to be developed at the suppressor grid of V3. The negative potential (approximately -2.1 volts) is obtained from the C9-R15 junction. The cathode potential of V3 is about +4.6 volts as a result of cathode voltage developed across R17 and C10. This further increases the suppressor-cathode negative potential.

Under these conditions (with the screen effectively controlling the tube current) the plate current is greatly reduced because the plate voltage has less influences on the electrons leaving the cathode than the screen. Many of the electrons, which do succeed in passing through the screen grid (and would normally add to the plate current) are repelled by the negative suppressor back to the screen grid. By selection of the proper relationship of plate and screen voltages, the respective currents will be approximately equal.

The negative suppressor potential produces a negative resistance effect in the tube. This is true because above a certain critical screen potential any further increases in suppressor voltage in a positive direction allows more electrons to be passed through the screen to the plate, resulting in a decrease in screen current. Since the screen current decreases, the amount of the suppressor bias produced by C9, R15, and R16 is correspondingly reduced. The reduction in the suppressor bias is equivalent to a positive-going suppressor voltage, which allows a further increase in the plate current and a further decrease in the screen current. Thus, the action is regenerative, and a small change in suppressor potential produces a greater effect on plate current than will a corresponding change in screen potential. The decrease of screen current with increase of suppressor volage (in a positive direction) indicates that the transconductance between the suppressor grid voltage and the screen current is negative.

When the circuit is oscillating, the alternating component of the screen tank comprising L1, C12, and C13 is applied simultaneously to the screen and suppressor grids. When the voltage is made more positive to ground by the tank oscillations, screen current is reduced as discussed. Conversely, when the tank voltage is going less positive, screen circuit is increased. The screen current therefore varies at the oscillator frequency. The screen output to the oscillator tank supplies the tuned circuit losses.

The oscillator frequency is 200 kc. A portion of the signal developed at the suppressor grid (at the R15-R16 junction) is fed through C1 to the suppressor grid of the first modulator, V2. It should be recalled that the video input is applied from V1 to the control grid of V2. Thus, the V2 output across C4, L2, and C5 consists of a 200-kc carrier frequency, which is amplitude modulated by the video pulsed input signal.

Modulator Amplifier

The modulator amplifier and cathode follower are combined in tube V4 (fig. 3-2). The modulated output of V2 is coupled by C7 to the control grid of V4A. This stage functions as a conventional class A amplifier and supplies its output through C15 to the grid of V4B. This section functions as a cathode follower whose output is developed across R23. V4B matches the impedance output of the modulator amplifier to the input impedance of the stages that follow.

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Horizontal and Vertical Modulators and Amplifiers

The horizontal and vertical modulators (V5, V6, V7 and V8 of figure 3-3) receive the modulated video input from the modulator cathode follower, V4B (fig. 3-2). The input to the modulators (fig. 3-3) is coupled through separate capacitors (C18, C23, C24, and C31) to the suppressor grid of its respective modulator. Thus, the same signal is applied to each of the modulators.

The horizontal modulator stages, V5 and V6, and the vertical modulators, V7 and V8, are connected in push-pull. V5 and V6 (the horizontal modulators) operate into separate tuned tanks. The plates are connected to the +150-volt supply through R55 and R56, respectively. V7 and V8 are the vertical modulators that also operate into separate tuned plate impedances, having their plates connected to the +150-volt supply through R45 and R46, respectively.

To produce a single pattern on the indicator screen, each of the modulator tubes (horizontal and vertical) must be driven so that their video outputs are balanced from saturation to cutoff of the modulator stages. By supplying the d-c potentials to all of the modulator stages from the same regulated source, and by the control of the video and no video input condition at the horizontal and vertical modulators from the first modulator (V2 of figure 3-2), each of the modulator stages (fig. 3-3) supplies video to the indicator during the same period.

The d-c potentials applied to the elements of each of the modulator tubes are approximately the same. Therefore, an analysis of the relationship of the potentials of V5 will be sufficient to show the relationship of the potentials that exists for all the modulator tubes.

The cathode of V5 connects through R44 and R43 to a -150-volt source. The V5 plate is connected to the +150-volt supply as discussed. Thus, V5 is connected between the -150- and +150-volt supplies. R43 and R44 are a part of a voltage divider, which develops a potential at the junction of the two resistors of about +0.7 volts. This potential is applied to the suppressor grid. The cathode d-c potential is more positive to ground (+3.5 volts) as a result of the addition of the voltage across R44. Thus, a negative bias exists between the cathodes and the suppressor grids of the modulators. The bias on the modulators is not sufficient to cause cutoff, and each of the modulator tubes is normally conducting.

Horizontal modulator control, R31, and vertical modulator control, R38, serve as balance controls for their respective modulators. By moving the arm of R31 clockwise (upward) the amount of the positive voltage at the V5 cathode will be decreased, while the positive potential at the cathode of V6 will be increased. Thus, the bias of V6 is increased and its gain is decreased.

Moving the R31 arm counterclockwise (downward) produces just the opposite effect to cause the gain of V5 to be decreased. Vertical control, R38, functions in the same manner. The action compensates for variations in the d-c resistance of the two tubes of the separate modulators. The vertical control is adjusted so that the output of both tubes is equal (balanced).

In addition to the modulated video input (which is fed to the suppressor grids of the vertical and horizontal modulators), a second input at the control grids is applied from an alternator. The rotor of the alternator, G1 (fig. 3-4) is driven by the shaft connected to the antenna drive motor. (See chapter 2, figure 8).

As the rotor turns (fig. 3-4), voltages are induced into the stator windings ("a" and "b"), which bear a 90° phase difference. The voltage developed by the "a" winding is applied to the grids of the horizontal modulators, V5 and V6 (fig. 3-3). R105 (fig. 3-4) provides a method of balancing the grid input to the separate modulator amplifiers. Likewise, voltage applied to the vertical modulators is fed from the "b" winding of the alternator. R116 acts as the grid balancing resistor for the input to V7 and V8.

The phase difference between the horizontal and vertical input signals at the alternator should be 90°. This relationship is necessary to produce the circular pattern on the cathode-ray tube of the indicator when no video input is being received.

A phase shifting network, comprising R102 and C58, is connected across the horizontal modulator input. R102 is variable so that the phase of the horizontal input can be adjusted to be exactly 90° relative to the vertical input.

The horizontal and vertical gain balance controls, R96 and R107 (fig. 3-3) respectively, function to control instantaneous changes in tube resistance (and tube conduction) of each of the modulator tubes. The action is necessary to prevent the introduction of distortion, which

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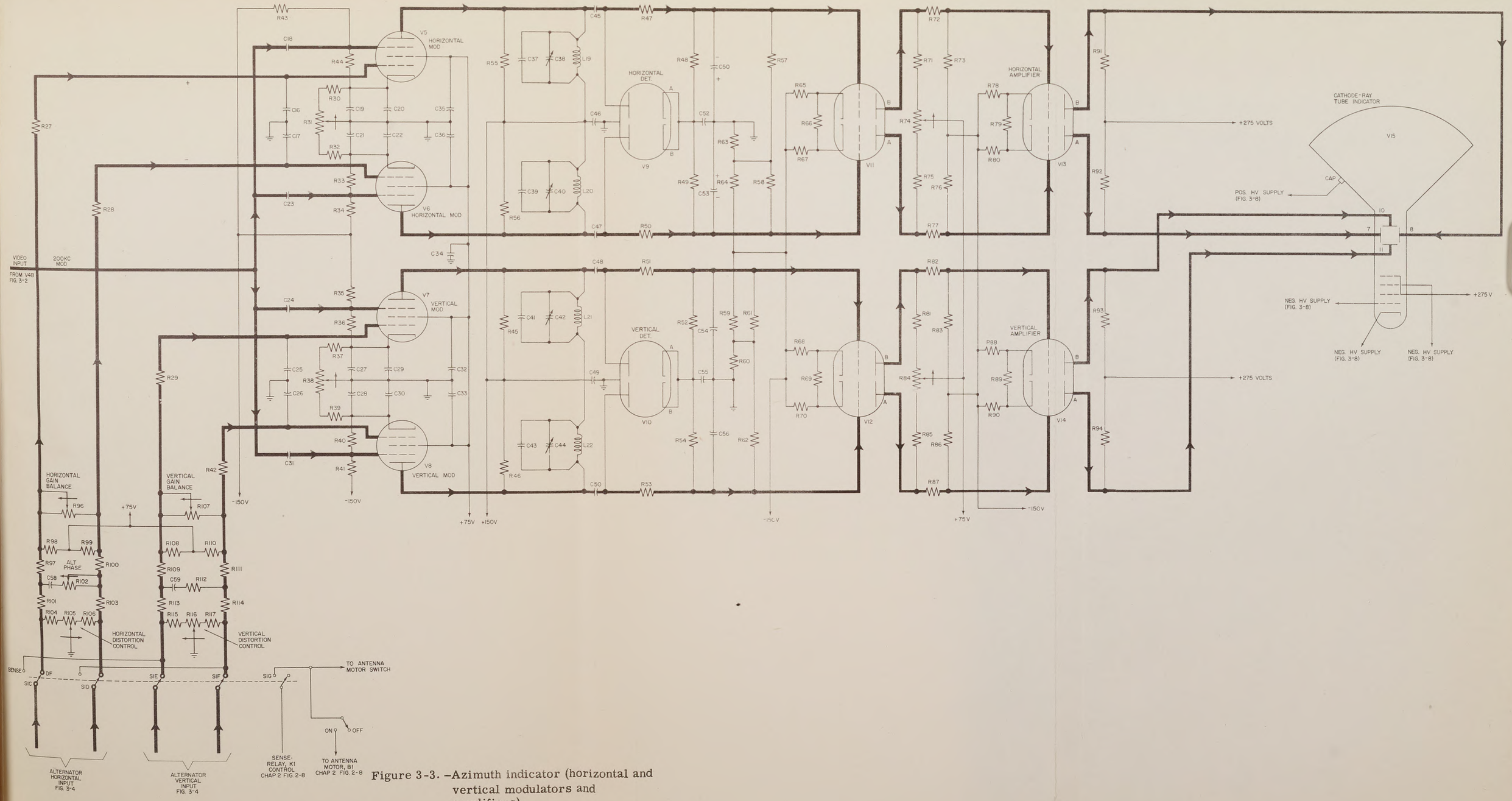

Figure 3-3. –Azimuth indicator (horizontal and vertical modulators and amplifiers).

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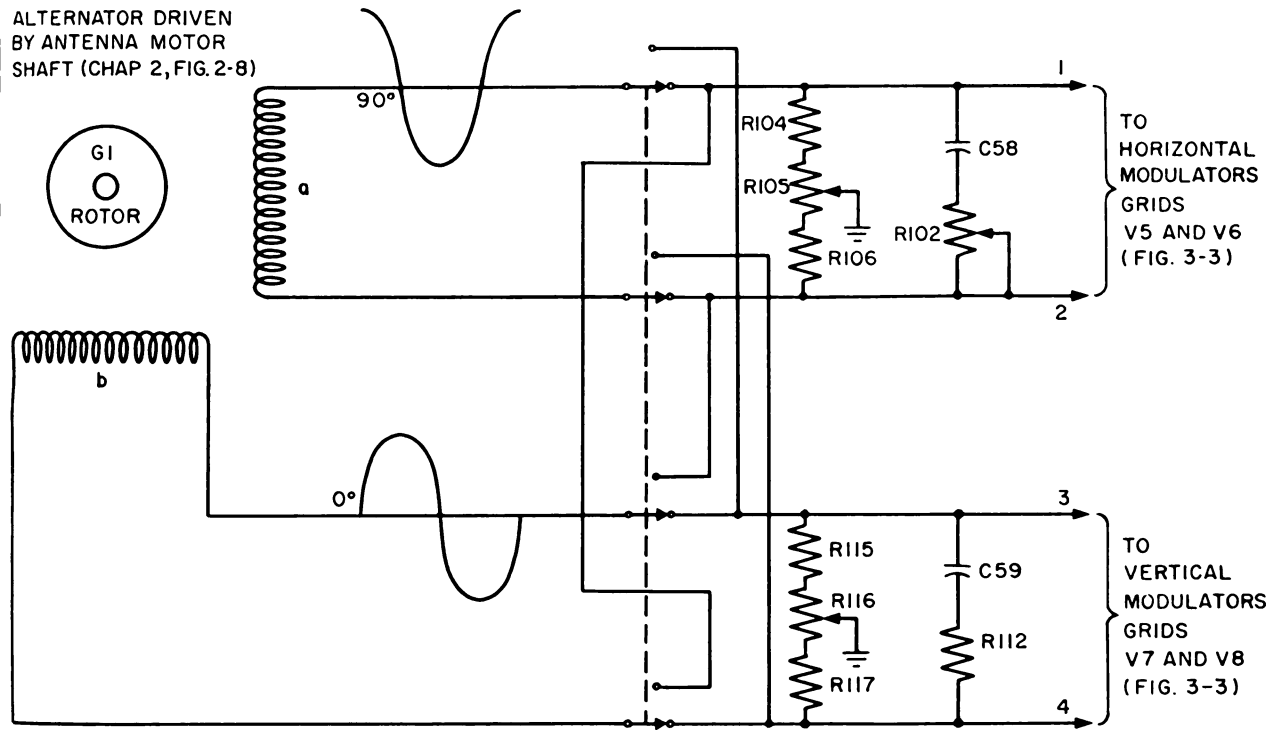

Figure 3-4. –Alternator and modulator input circuit.

would otherwise result in nonsymmetrical output amplitude signals from the horizontal (or vertical) modulator tubes.

The simplified diagram of figure 3-5 should be helpful in understanding the circuit operation. Assume that R31 has been set so that the conduction of both sections of the horizontal modulator is equal. The corresponding fixed resistors in each section of the bridge are of equal value so that the current through path abc is equal to the current through path adc, if the V5 and V6 grid-cathode resistances are equal. In this condition, the voltage bd (across R96) is zero.

If the grid-cathode resistance of V6 decreases, the current through path adc increases. Any small increase in this current causes a large change in the voltage drop across R99, relative to the voltage change across the other smaller resistances in this bridge section. As a result, the voltage cd increases, while the voltage da decreases.

Since the voltage division abc has remained the same, point b is now positive to point d. The resulting current through R96 causes a less positive voltage toward the V6 grid, and a more positive potential at the grid of V5. Thus, the conduction of V5 increases, while that of V6 decreases. The voltage across R98 increases, and the voltage across R99 decreases. This action prevails until the potential across R96 is again zero.

Actually, any tendency of one tube to conduct a heavier current than the other receives an immediate counteraction. Thus, a closer observation of the circuit will show that the potential across R96 never rises above a fraction of a volt so long as both tubes are operating within their conduction ranges.

If section abc of the bridge attempts to produce a predominating current, the reverse action to that discussed will occur. R96 may be adjusted to balance the conduction of the tubes when larger than normal current variation occurs, as may be the case when either or both of the tubes is replaced.

The output of the horizontal modulators, V5 and V6 (fig. 3-3) is developed across the tank circuits comprising L19, C37, and C38, and L20, C39, and C40, respectively. The phase of the horizontal output leads the phase of the vertical output by 90° as a result of the relative phase of the alternator input to the vertical and horizontal modulators. The horizontal output is

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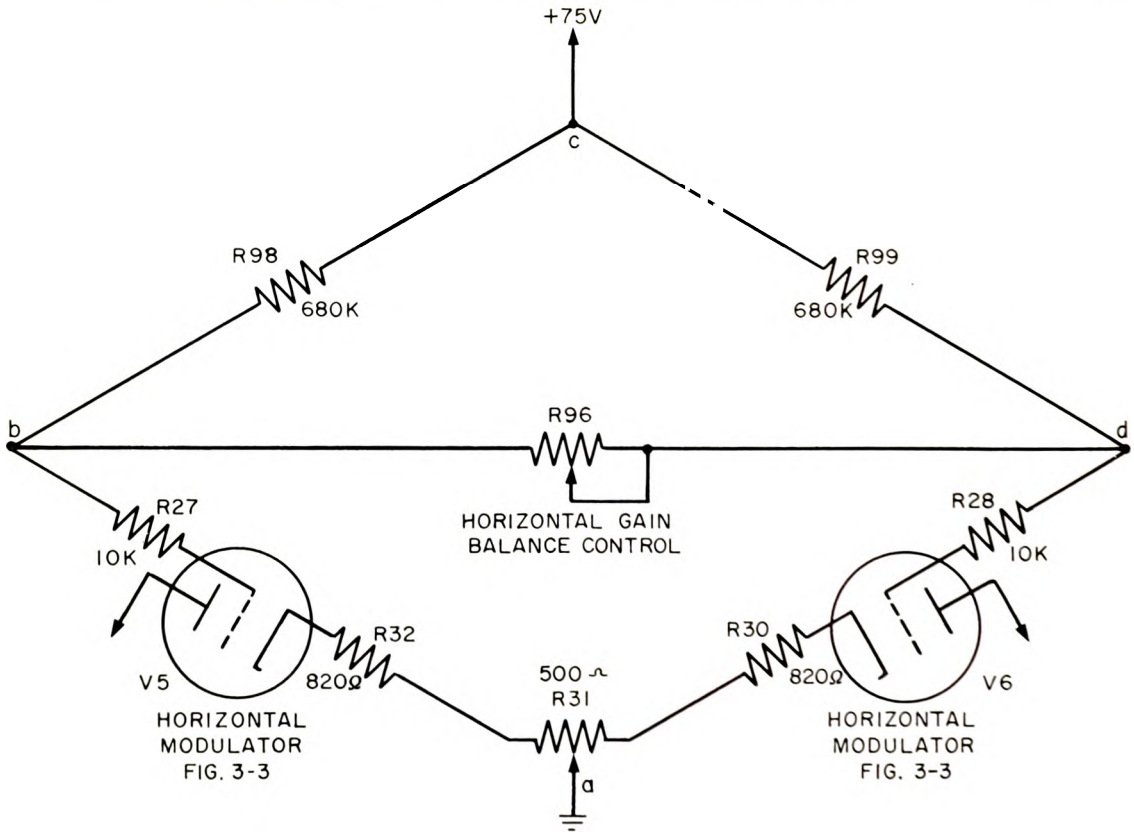

Figure 3-5. –Horizontal gain balance control.

coupled by C45 and C47, respectively, to separate horizontal rectifiers, V9A and V9B. The vertical output of V7 and V8 (developed across L21, C41, and C42, and L22, C43, and C44) is coupled by C48 and C50, respectively, to the separate vertical rectifiers, V10A and V10B.

The horizontal and vertical rectifiers, V9 and V10, function to convert the modulated r-f input from the horizontal and vertical modulators into a video pulsating direct current output. It should be remembered that the voltage at the V5-V6 plate tanks is 90° out-of-phase with the signal developed by the V7-V8 plate tanks.

The plate tanks of the horizontal and vertical modulators do not respond to the video and alternator frequencies since these frequencies are below the resonant frequency (200 kc) of the tanks. Thus, the signal fed from the plate tanks of the modulators to the detectors, V9 and V10, is a modulated radio frequency. Since each of the detectors combined in V9 and V10 functions the same, only the detector circuit of V9A will be discussed.

Note that the alternator 15-cps input to the horizontal modulators, V5 and V6, from terminals 1 and 2 (fig. 3-4) will be negative toward one of the grids of the horizontal modulators (fig. 3-3) while positive toward the other grid. In this manner, the grids of V5 and V6 are always fed 180° out-of-phase as a result of the alternator input.

It should be recalled that the 200-kc a-m signal from V4B (fig. 3-2) is fed in phase (at the suppressor grids) to each of the modulator stages, V5 through V8 (fig. 3-3). Thus, the modulator receives both the video and alternator signals, as discussed. (Video as used in this discussion refers to the component of the signal voltage, which is received at the RDF antenna, and not to the signal frequency.)

Assume an instantaneous input polarity from the alternator to the V5 and V6 grids, as shown. Due to the more positive V5 grid, V5 will produce a greater driving signal to the V5 plate tank than will be produced for the V6 plate tank. As a result of this action, a greater magnitude

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signal voltage will be applied to the V9A horizontal rectifier through C45 than will be fed to the V9B rectifier through C47.

If an unmodulated 200-kc signal (containing no video or alternator voltage) were applied to V9A and V9B, both diode sections would produce equal voltages negative to ground across C50 and C53, respectively. The negative voltage results from the discharge of C45 and C47, respectively, during negative alternations of the plate tank input. Since V9A and V9B will produce equal amplitude outputs negative to ground across C50 and C53 for an unmodulated 200-kc input, the algebraic sum of the voltages will be zero.

When the 200-kc modulated signals from V5 and V6 are developed across the plate output tanks, one output will be of greater magnitude than the other due to the positive alternator input at one control grid (the grid of V5 in this case), and a more negative input to the V6 control grid. Thus, the negative voltage to ground across C50 will be of greater magnitude than the negative voltage to ground across C53. The detector output voltage at any instant is the algebraic sum of the C50 and C53 voltages.

Note that the negative voltage to the control grid of horizontal amplifier, V11B (from the C50-R57 junction) is more negative, while the voltage to the V11A control grid (from the C53-R58 junction) is less negative. Thus, the two grids of the horizontal amplfiers, V11B and V11A, are fed 180° out-of-phase.

The video signal applied from the vertical detector, V10A, to the vertical amplifier, V12B (in the same manner as described for V9A) is 90° out-of-phase with the signal from V9A to V11B. The 90° phase difference in the vertical and horizontal signals must exist, and must be preserved, in order to produce the various indicator patterns that will be discussed later. The 90° phase difference in the vertical and horizontal signals was introduced as a result of the alternator voltage input at the horizontal and vertical modulators, V5 through V8, as discussed.

The horizontal and vertical amplifiers, V11 through V14 (fig. 3-3) increase the video signal amplitude, which is to be applied to the respective deflection plates of the cathode-ray tube in the indicator discussed later.

The effect of a phase shift at the input to a conventional electron-tube amplifier caused by the series combination of the coupling capacitor and grid resistor is seldom considered, although the phase of the input signal may be shifted by the input components in some cases more than 60°. Changing the value of the grid components will, of course, change the amount of the phase shifting at the amplifier grid. However, a zero phase shift across a reactive component with respect to a series connected resistive component cannot be accomplished. For this reason, direct-coupled amplifiers are employed in the horizontal and vertical amplifiers, V11 through V14, to eliminate the need for coupling capacitors.

In addition to the phase-shifting action just discussed, it should be noted that if a capacitor were used to couple the video signal from one of the amplifier stages to the other, the charge time of the coupling capacitor would not only delay presentation of the pattern, but would also cause the indicator pattern to move either upward or downward on the screen as the capacitor is charged and discharged by the incoming video signal. Also, the size of the capacitor would have to be relatively large in order to offer low reactance to the various components of the video signal.

V11 and V13 (horizontal amplifiers) are series connected between the -150- and +275-volt supplies. The cathode of V11B is connected to the -150-volt supply and maintained at -80 volts. This potential is derived by the conduction of V11B from the -150-volt source, through R65 and the plate-to-cathode resistance of V11B, and through R71 and R74 to the +75-volt supply. The grid potential of V11B is held at -55 volts. This potential is obtained from a separate voltage divider (comprising R64 and R63), which is also connected from the -150-volt supply to ground. Thus, during quiescence, the grid potential (-55 volts) of V11B is positive with respect to the cathode (-80 volts), and grid current flows. Plate current also flows through R71 and R74.

The flow of the V11B grid current through R57 holds the grid voltage within a fraction of a volt positive to the cathode. This is true since a very small portion of the total -150 volts applied will exist from the cathode to the grid, due to the low grid-cathode resistance, relative to the large resistance values of R65, R57, and R63 in series.

The positive-going video signal from V9A increases the current through V11B and causes the V11B plate voltage to decrease. Note that the potential at the R71-R72 junction is the same as the V11B plate potential. R71 and R72 are

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a part of a voltage divider comprising R73, R72, R71, and R74, which is connected from the -150-volt supply to the +75-volt supply. Since the potential at the R71-R72 junction has decreased to ground, the R72-R73 junction potential also decreases. This decrease in potential (at the R72-R73 junction) is applied between the control grid and cathode of the horizontal amplifier, V13B, via cathode resistor, R78.

As a result of a less positive-going input to the V13B grid, V13B conducts a smaller current through R91 to the +275-volt supply. The potential rise at the V13B plate, which is the amplified video signal, is directly applied to the pin 8 horizontal deflection plate of the cathode-ray indicator tube, V15.

At any given time, the grid input signal to the V11A and V13A sections of the horizontal amplifiers is 180° out-of-phase with the signal applied to the V11B and V13B amplifiers. Thus, the V13A output to the pin 7 horizontal plate of V15 is less positive during the same period that V13B is supplying a more positive potential to the pin 8 horizontal plate of V15. The simultaneous application of the video signal to both horizontal deflection plates of V15 increases the control of the horizontal plates on the electron stream.

The vertical amplifiers, V12 and V14, function the same as the horizontal amplifiers, V11 and V13. The output of V14 is a video signal to the vertical plates (pins 10 and 11) of V15, which is 90° out-of-phase with the video applied to the horizontal deflection plates (pins 7 and 8), as discussed. The action of the horizontal and vertical voltages at the cathode-ray tube, V15, to produce the various indicator patterns will be considered presently.

DEVELOPMENT OF INDICATOR PATTERNS

CIRCULAR PATTERN

If no video input is applied to the first modulator (V2 of figure 3-2) from the video differential amplifier, V1, V2 will conduct continuously, and only the 200-kc oscillator output of V3 will be fed through V2 and V4 to the modulators (V5 through V8 of figure 3-3). In the modulator stages, V5 through V8, the 200-kc radio frequency generated by V3 (fig. 3-2) will be modulated by the vertical and horizontal 15-cycle alternator inputs (fig. 3-3), which are 90° out-of-phase. The modulators, V5 through V8, conduct continuously, and the resulting output to the vertical and horizontal deflection plates of V15 (after passing through the detector and amplifier stages) produces vertical and horizontal sine wave voltages, respectively, which are 90° out-of-phase. Thus, the circular pattern shown in figure 3-6, A, is produced, representing zero video signal input at the antenna.

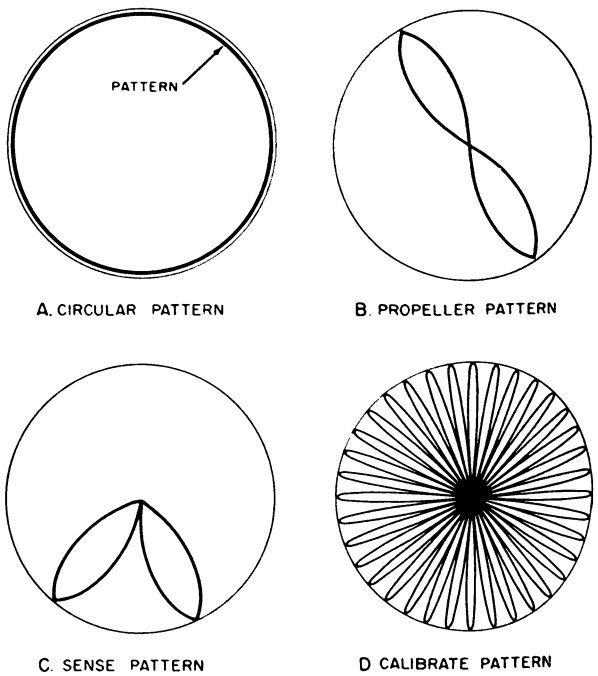

Figure 3-6. – Indicator patterns.

PROPELLER-SHAPED PATTERN

Signals received at the antenna are 15 cps modulated by the antenna rotation. The modulation introduced on the incoming signal is effective in controlling the output of the video differential amplifier, V1 (fig. 3-2), as discussed earlier.

The video signal at the V1B grid is negative going. When the V1B input is maximum negative (corresponding to maximum received signal at the antenna) the effect is to increase the V1A plate current through R6, which cuts off the first modulator stage, V2, and no 200-kc carrier from V3 can pass to the modulators, V5 through V8 (fig. 3-3). When the V1B input is minimum (corresponding to the nulls of the received signal), the V1A plate current through R6 is decreased and the first modulator, V2, is

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conducting. A video signal will be passed through the modulators, detector, and amplifiers of figure 3-3 only when the first modulator, V2, is conducting, or when the nulls of the received signals are being applied to the V1B grid.

The period of conduction of the first modulator, V2, is small as compared to the total duration of a single pulse of the video signal. Thus, the signal applied to the vertical and horizontal deflection plates of V15 (fig. 3-3) is a pulse, which corresponds in time to the reception of the nulls of the received signal.

The phase relationship of the vertical and horizontal pulses is always 90°, as discussed. If the vertical and horizontal voltages were complete sine waves, a circular pattern would be produced for the 90° phase relationship. However, since the vertical and horizontal inputs are pulses, the pattern displayed on the indicator (V15) is in the form of a propeller, the tips of which point in the direction of the nulls of the received video. An azimuth scale which encircles the cathode-ray tube screen (chapter 2, figure 1) permits the operator to read, in degrees, the position of the two nulls of the received signal.

SENSE PATTERN

The circuit of the antenna in D/F operation is as shown in chapter 2, figure 9, A. The two sections of the adcock array are in exact balance during D/F operation, as discussed.

When the calibrate-sense switch on the front panel of the indicator (chapter 2, figure 1) is held in the SENSE position, a sense relay, K1 (chapter 2, figure 8) energizes, allowing its contacts to alter the antenna electrical connections to that shown in figure 2-9, B. The effect is to introduce a phase lag in one section of the array, while simultaneously introducing a phase lead in the other section. This action causes the nulls to appear at a different position in the rotation of the antenna, and at a different position on the indicator screen.

In addition to the action just discussed, when the sense-D/F switch is depressed, sections S1C, S1D, S1E, and S1F of S1 (fig. 3-3) interchange sweep voltages to the balance modulators, V5 through V8. This causes the pattern to be shifted 90° on the indicator screen.

The combined effects of the sense-D/F switch, S1, and the sense relay, K1 (chapter 2, figure 8) causes the rotation of the pattern by 90° and a shift in the position of the nulls, so that the distorted figure 8 (sense pattern of figure 3-6, C) is formed. The apex of the distorted figure 8 points in the direction of the received signal.

CALIBRATION PATTERN

The calibration display is a 36-lobe "daisy" pattern (fig. 3-6, D) displayed on the cathode-ray tube, V15. The 36-lobe pattern is used to calibrate the indicator.

A tone wheel, E1 (fig. 3-7), which is driven by the antenna motor, B1 (chapter 2, figure 8), affects the core material of L5A (fig. 3-7) to generate the calibration voltage in the circuit comprising L5A, C5, and the primary of T1. This voltage is developed at the T1 secondary across the calibration gain control, R173.

The calibration phase splitter, V23A, receives the calibration signal input from the arm of R173. The plate and cathode resistances of V23A are of equal value to provide equal amplitude outputs. R175, C77, R174, and the tube resistance of V23A form a phase shifting network for the plate output of V23A. The amount of the phase shift is determined by the position of the arm of R174. The cathode potential of V23A (which is applied to the arm of R174) is degenerative, since it is approximately 180° out-of-phase with the plate output. The degenerative action is reduced by a smaller regenerative action due to the charge of C77 through R177 on the negative half cycle of the input, and the discharge of C77 through V23A during positive portions of the input. The resulting phase of the signal obtained from the combination of the plate and cathode voltages is applied to the grid of amplifier, V23B.

V23B is a conventional amplifier stage. When sections S1A and S1B of the sense-D/F cal switch are operated to the CAL position (as shown), the V23B output is coupled by C79 to the control grid of the first modulator stage (V2 of figure 3-2). Thus, during calibration the V23B output causes the first modulator, V2, to produce an output pulse 36 times for each revolution of the antenna. The resulting signal to the vertical and horizontal deflection plates of the cathode-ray tube, V15 (fig. 3-3), after passing through the modulators, detector, and amplifiers, produces 36 lobes on the indicator screen during the same period normally required for the two-lobe (single-propeller) pattern to be produced. The 36-lobe pattern is shown in figure 3-6, D.

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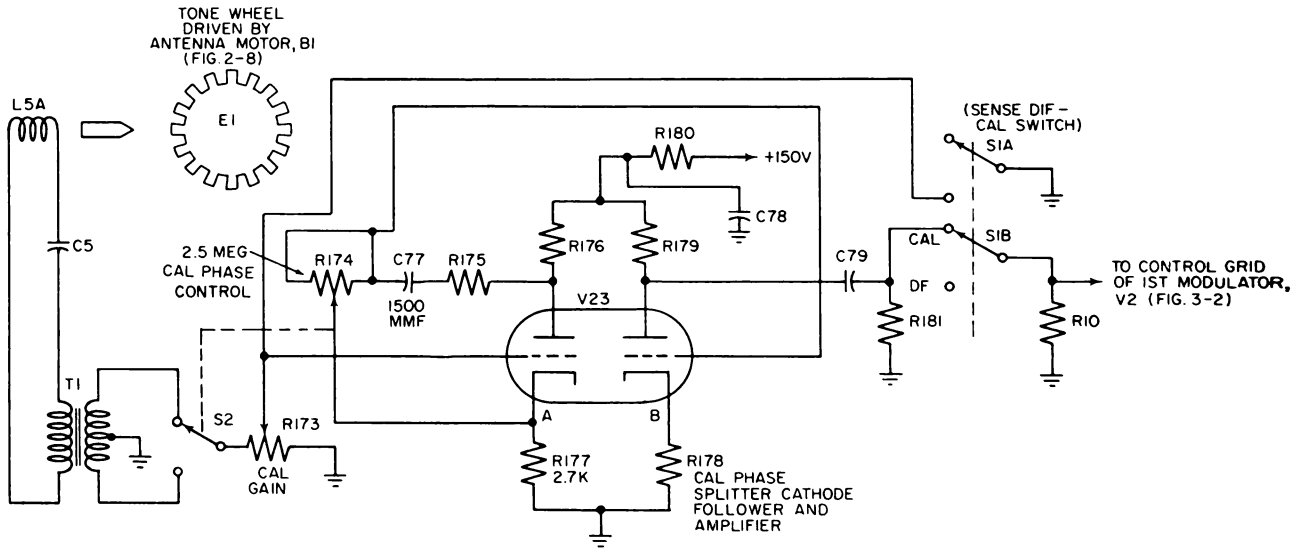

Figure 3-7. –Calibration voltage input.

The phase of the V23B output (fig. 3-7) can be shifted by rotating the cal-phase control, R174. This action simultaneously changes the position of all the lobes of the calibrate pattern (fig. 3-6, D). In this manner, the calibration necessary for determining the direction of a received signal can be accomplished by moving the tips of the 36-lobe pattern (spaced every 10 degrees) until 4 of the tips correspond with 0, 90, 180, and 270 degrees on the azimuth indicator.

The cal-gain control, R173 (fig. 3-7) is used to set the output voltage level at V23B. This action will adjust the amplitude of the signal pulses at the deflection plates of V15 (fig. 3-3). The control is properly adjusted when the spokes of the 36-lobe pattern converge in a spot near the center of the cathode-ray tube screen.

When the sense-D/F-cal switch, S1 (fig. 3-7) is placed in the D/F position, the grid of V23A is directly grounded. This prevents the amplification of the calibrate signal in V23, and eliminates the possibility of allowing an amplified calibrate signal to be capacitively coupled to the CRT deflection plates through the open contacts of S1.

HIGH-VOLTAGE POWER SUPPLY

The high voltage for the cathode-ray tube, V15 (fig. 3-8) is supplied by rectifying an r-f voltage generated by an oscillator, V16. The r-f oscillator output is transformer coupled from the 2-3 winding of T2 to the high-voltage winding, to the directly heated cathode of V17, and to the plate of V18. Feedback for the oscillator, V16, is developed by a section of T2, which is tied to the V16 control grid.

The oscillator output voltage is stepped up across the high-voltage secondary of T2. V17 produces a negative output to ground across the filter capacitors, C80 and C81. V18 produces a positive voltage across its filter capacitors, C68 and C69. Small values of capacitance are used in the negative and positive supplies since the ripple frequency is high (about 265 kc), and the current drain from the supply by the cathode-ray tube is small.

R132 through R139 make up the bleeder (voltage divider) resistance across the negative supply filter. The full value of the negative voltage (from the R139 through R151 junction) is applied to the control grid of the CRT. The cathode potential is obtained from a less negative point (between R137 and R138) on the divider.

The pin 9 (1st anode) voltage for V15 is obtained from the arm of R131. This resistor is connected across the +275-volt supply (not shown). The full value of the positive supply output, V18, is applied from the R145 through R146 junction to the cap (2nd anode) of the CRT.

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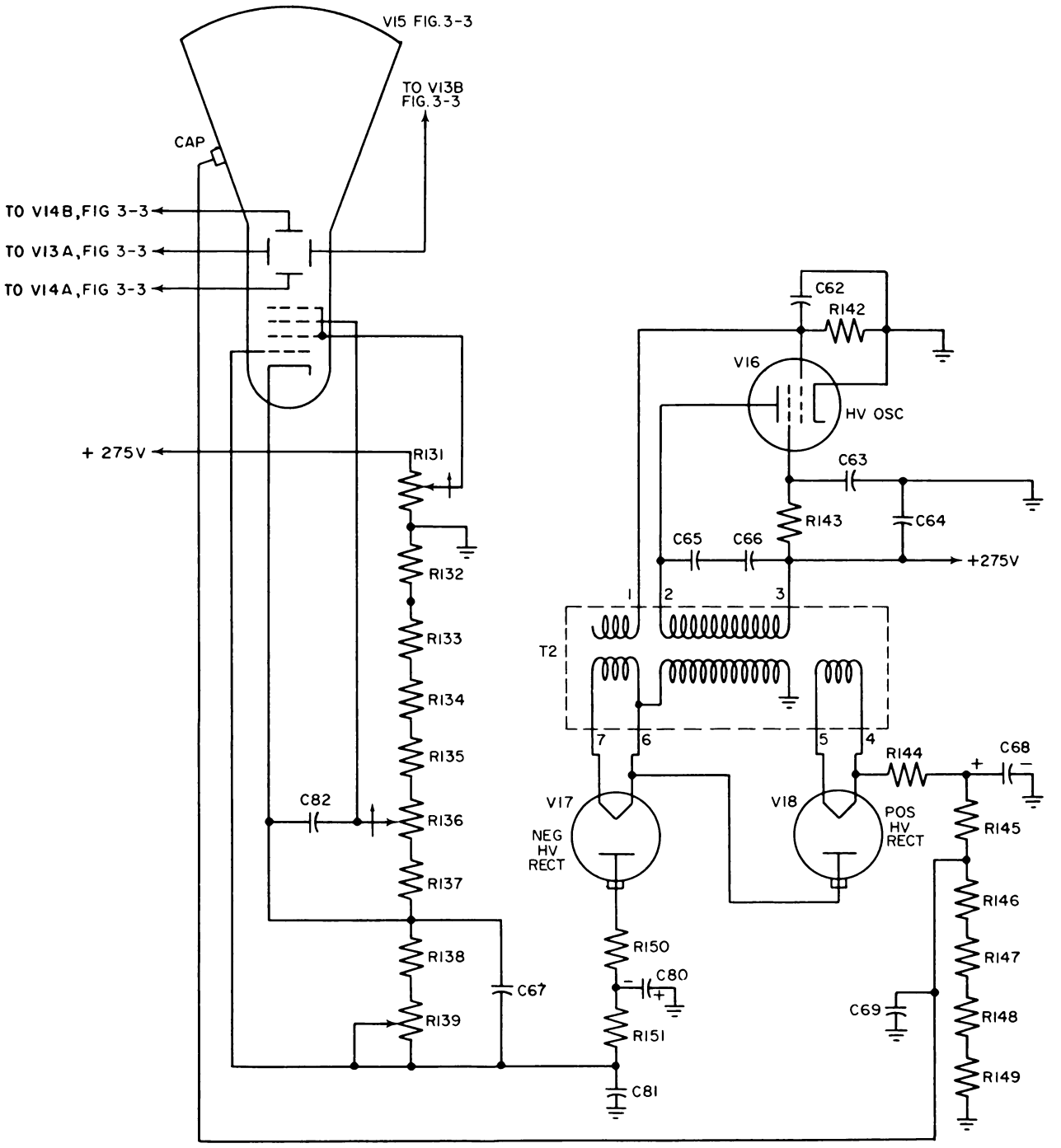

Figure 3-8. –High-voltage power supply.

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QUIZ

1. The primary purpose of the i-f video audio chassis (fig. 3-1) is to perform what function?
2. The input to the third i-f amplifier stage V29 (fig. 3-1) is simultaneously applied to what other stage?
3. Signals applied at the V32B plate (fig. 3-1) must exceed what value of voltage in the positive direction in order to produce an AGC output current?
4. The automatic gain control voltage for the r-f stages in the r-f head (chapter 2, figure 10) is produced by what stage in figure 3-1?
5. The action of both sections of V32 (fig. 3-1) to produce the AGC voltage is referred to as what type action?
6. The effect of the AGC voltage to decrease the gain of the r-f stages (fig. 3-1) can be controlled by adjusting the position of the arm of what control?
7. What is the purpose of C127 (fig. 3-1)?
8. That portion of the audio output developed across R203 (fig. 3-1) is also available across what other component?
9. When a carrier signal is detected by the signal rectifier, V30A (fig. 3-1) a negative (or less positive) potential is fed to the grid of what audio stage?
10. When the local remote switch, S2 (fig. 3-1) is placed in the LOCAL position, +105 volts is directly applied to the V34A cathode for what purpose?
11. Z1 in the plate circuit of V33 (fig. 3-1) allows maximum passage of frequencies between what range?
12. When the tuning system (discussed in chapter 2) is tuning the r-f head towards a lower frequency relay K8 (chapter 2, figure 18) is deenergized to cause what effect on the audio (V35) output (fig. 3-1)?
13. What chassis of the AN/URD-4 (chapter 2, figure 1) provides the means by which the detected video signals from the video-audio chassis (fig. 3-1) are amplified mixed with the sweep voltages originating at the antenna, and presented to the cathode-ray tube for display?
14. In the absence of an input signal, the two sections of V1 (fig. 3-2) may be made to conduct equal currents by adjusting what resistor?
15. The 200-kc carrier oscillator (V3 of figure 3-2) is connected as what type circuit?
16. Horizontal modulator control, R31, and vertical modulator control, R38 (fig. 3-3) serve as what type controls for their respective modulators?
17. R102 (fig. 3-3) is variable so that the phase relationship of the horizontal and vertical inputs can be adjusted to exactly what value?
18. V11 and V13 (horizontal amplifiers of figure 3-3) are series connected between what two power supplies?
19. In the modulator stages (V5 through V8 of figure 3-3) the 200-kc oscillator (V3) r-f input is modulated by what signals?
20. The high voltage for the cathode-ray tube, V15 (fig. 3-8) is obtained by what means?

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CHAPTER 4

OPERATING PRINCIPLES OF A REPRESENTATIVE RADIO RECEIVER

INTRODUCTION

Modern Navy receivers use rugged components, which utilize circuits that make them easy to operate and maintain. They are capable of receiving several types of signals and can be tuned accurately.

The circuits of the AN/SRR-11, 12, and 13 receivers described in this chapter represent equipments widely used in the Navy. You should study them carefully.

The AN/SRR-11, 12, and 13 receivers consist of several major assemblies, each of which is physically independent of the other. Individual subassemblies are mounted with each assembly, and they plug into an appropriate socket of their major assemblies. When it is necessary to make repairs and time is an important factor, a defective assembly or subassembly can be replaced quickly with a spare one.

OPERATIONAL USE OF RADIO

Radio frequencies extend from 10 kc to 300,000 mc and are divided into various bands for convenience. The bands used for military purposes are shown in table 4-1.

Each of the bands have certain transmission characteristics, which are peculiar to that band. Occasionally the characteristics within a band may differ, but on the average they are consistent (see table 4-2).

The terms, long range, medium range, and short range, correspond to 1500 miles, 200 to 1500 miles, and under 200 miles, respectively.

Very Low-Frequency Band

The very low-frequency band (from 10 to 30 kc) is not covered by shipboard transmitters. The transmitting antennas required for these frequencies are too long to use aboard ship. However, shore stations, such as NAVCOMMSTA Washington (NSS) employ the v-1-f band for fleet broadcast. Powerful v-1-f stations with their huge antennas are capable of transmitting signals through magnetic storms that will ordinarily blank out the higher frequency bands.

Table 4-1. — Frequency Spectrum

Band	Frequency range
Very low frequency (vlf)	10 to 30 kc
Low frequency (lf)	30 to 300 kc
Medium frequency (mf)	300 to 3,000 kc
High frequency (hf)	3 to 30 mc
Very high frequency (vhf)	30 to 300 mc
Ultra high frequency (uhf)	300 to 3,000 mc
Super high frequency (shf)	3,000 to 30,000 mc
Extremely high frequency (ehf)	30,000 to 300,000 mc

Low-Frequency Band

The direction-finding equipment used by the Navy generally employs frequencies in the 30- to 300-kc range. Reliable long-range communication can usually be obtained in the l-f band with low-frequency equipments and antennas aboard ship. Frequencies in the l-f band (like those in the v-l-f band) do not depend on sky waves and provide stable communications with little variation from season to season.

Medium-Frequency Band

Transmission of frequencies in the 300- to 3000-kc range usually employs the use of ground waves even though sky wave transmission and reception of m-f waves are also possible. At the

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Table 4-2.—Frequency Transmission Characteristics.

Band	Range		Power required	Antenna length requirements
	Day	Night		
VLF	Long	Long	Very high	Very long
LF	Long	Long	Very high	Long
MF	Medium	Long	High to medium	Long
HF (3 to 10 mc)	Short	Medium to long	Medium	Medium
HF (10 to 30 mc)	Long	Short	Low	Short
VHF	Short	Short	Low	Very short

upper end of this band, the ionosphere has a great effect on the sky wave. If the correct frequency is used at the proper time, relatively large distances can be obtained by using this band. Propagation characteristics vary from season to season, and from night-time to daytime transmission.

The international distress frequency (500 kc) is located in this band. Commercial broadcast as well as some Navy stations employs frequencies in the m-f band. Modern Navy transmitters designed to operate in the m-f band are the AN/SRT-14, 15, and 16, which are discussed in chapters 5, 6, and 7 of this training course.

High-Frequency Band

The h-f band includes frequencies from 3 to 30 mc. The sky wave is used for long-range transmission in this band. Propagation characteristics of frequencies in this range also change with the time of day and the season of the year. The h-f band is divided into two parts (table 4-2) to show the characteristics of each part.

Transmitters such as the TBM, TBL, TBK, and TDE cover a part of the m-f and most of the h-f bands. The TBL and TDE transmitters also cover a portion of the l-f band. Although these transmitters are obsolete, they are still used aboard some ships. The AN/SRT-14, 15, and 16 also operate in the h-f band.

Very High-Frequency Band

The short antennas used in the 30- to 300-mc band require less space than lower frequency antennas and thus have an advantage for shipboard use. In addition, the power required for transmission in the v-h-f range is relatively low.

For frequencies greater than 30 mc, the surface wave component of the ground wave is rapidly attenuated through the dielectric constant of the earth. For this reason, the direct wave (that component of the ground wave traveling directly from the transmitting antenna to the receiving antenna) is the primary component of v-h-f propagation. The distance covered by a given transmission in this range is therefore limited by the distance to the horizon, or line-of-sight distance from the transmitter.

From this discussion, it is easy to see that several factors must be considered when a certain frequency is to be selected for transmission. The Electronics Technician has little to say about the choice of this frequency. However, a knowledge of the factors discussed should be helpful in determining why that particular frequency was chosen.

GENERAL DESCRIPTION

The AN/SRR-11, 12, 13 receivers are designed for Navy shipboard installation and cover the frequency range from 14 kc to 32 mc. The AN/SRR-11 operates at the low-frequency end of the frequencies covered, while the AN/SRR-12 and 13 cover the medium and high frequencies, respectively. Each of the receivers has its total frequency range divided into five bands, as shown in table 4-3. Continuous tuning within each band is accomplished by the tuning controls located on the front panel of the receiver. The AN/SRR-11 is capable of receiving c-w(A_1), m-c-w(A_2), and frequency-shift keyed

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Table 4-3.—Tuning Bands and Range of Each Band.

Band	Low-Freq. receiver AN/SRR-11	Medium-freq. receiver AN/SRR-12	High-freq. receiver AN/SRR-13
1	14 to 30 kc	.25 to 0.5 mc	2.0 to 4.0 mc
2	30 to 63 kc	0.5 to 1.0 mc	4.0 to 8.0 mc
3	63 to 133 kc	1.0 to 2.0 mc	8.0 to 16 mc
4	133 to 283 kc	2.0 to 4.0 mc	16 to 24 mc
5	283 to 600 kc	4.0 to 8.0 mc	24 to 32 mc

signals (F_1), while the AN/SRR-12 and 13 are equipped with circuits that permit voice (A_3) reception in addition to those mentioned.

The AN/SRR-11 is shown in figure 4-1, while the AN/SRR-12 and 13, whose external appearances are the same, are represented by a single picture in figure 4-2. The operating controls are mounted on the front panel of the respective receivers. We will refer to these controls, as they are named on the front panel, while discussing their associated circuits.

The circuits of the three receivers are the same in many ways. However, in some sections of each receiver the circuits are sufficiently different to warrant separate discussion. For this reason, the first portion of this chapter will be limited to a discussion of the AN/SRR-11, which, as forestated, is designated the low-frequency receiver. The latter portion will be devoted to the circuit differences contained in the AN/SRR-12 and 13, designated the medium- and high-frequency receivers, respectively.

BLOCK DIAGRAM OF THE AN/SRR-11

A block diagram of the AN/SRR-11 is shown in figure 4-3. We will use this figure to trace the signal path through the receiver.

Signals coupled from the antenna are amplified in two stages of r-f amplification. The first stage, the antenna preamplifier, couples its output to the second stage, the r-f amplifier. The bias, and therefore the gain of the amplifiers, is controlled by an r-f gain control connected in the two stages.

The r-f amplifier couples its output to the mixer. In addition to the signal received from the r-f amplifier, the mixer receives a signal from the local oscillator. The output of the mixer is determined by the band to which the receiver is tuned. On bands 1 and 4, the mixer output frequency is 60 kc. On bands 2, 3, and 5,

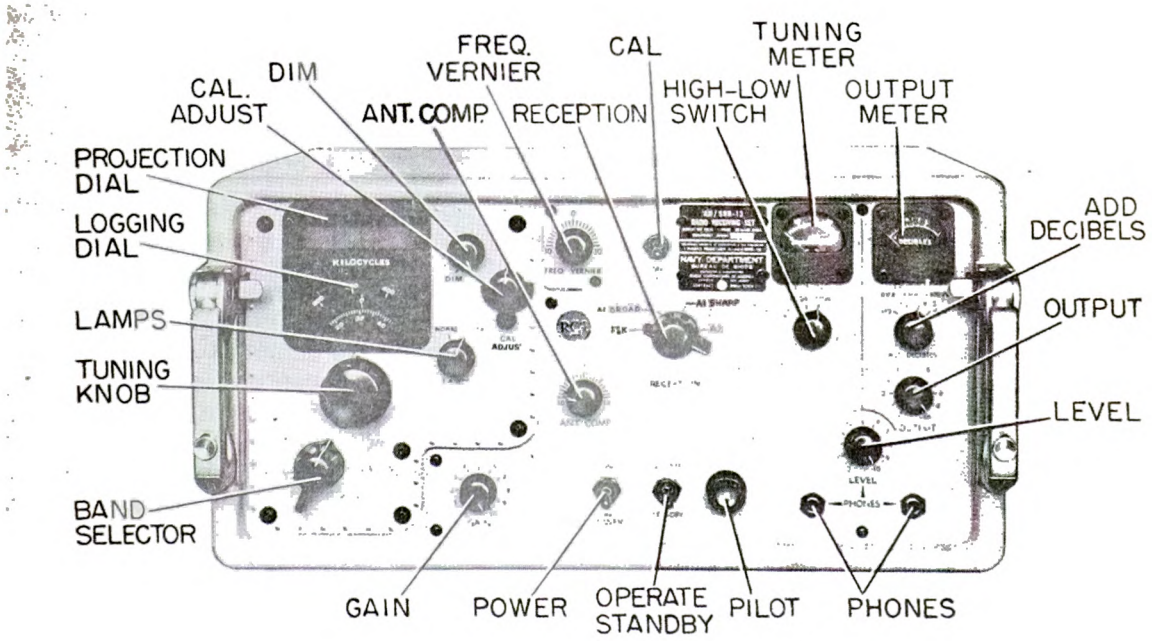

Figure 4-1. —Receiving cabinet and front panel of AN/SRR-11.

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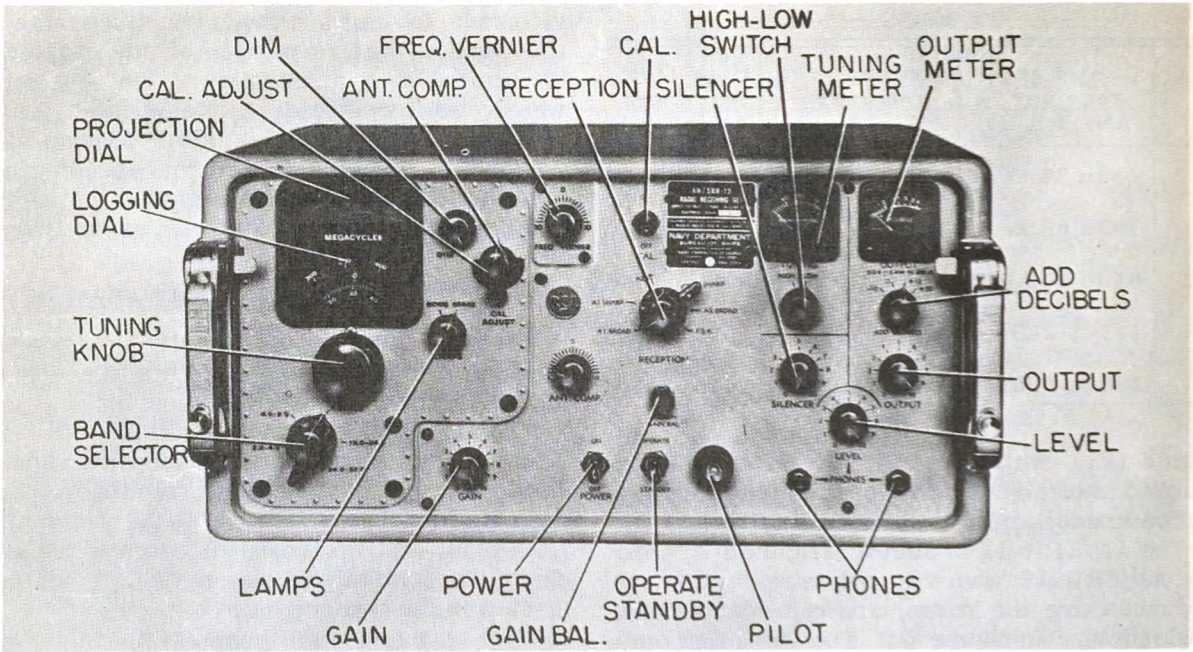

Figure 4-2. —Receiving cabinet and front panel controls of the AN/SRR-12 and 13.

the output frequency of the mixer is 200 kc. When the output frequency of the mixer is 60 kc (bands 1 and 4) the band switch, S1, is set so that the signal passes through the filter and the first i-f amplifier.

The first i-f amplifier of the first i-f assembly is a frequency converter for bands 1 and 4. On bands 2, 3, and 5, it is an i-f amplifier. It receives the 60-kc signal from the filter and mixes it with a 140-kc signal generated within the stage to produce a sum frequency of 200-kc output.

On bands 2, 3, and 5 the mixer output is already 200 kc and needs no further mixing. Therefore the first i-f assembly is bypassed entirely (S1 in position shown).

The output of the first i-f assembly, whether heterodyned or bypassed, is coupled to the second i-f assembly. The latter section consists of filters, three stages of 200-kc i-f amplification, a diode detector, a beat frequency oscillator and mixer, and a cathode follower.

The signal enters the second i-f assembly through either of two filters. The filter selected depends on the setting of the reception control switch, S2A, which is located on the receiver front panel (fig. 4-1). When the switch is set at FSK, A1 BROAD, or A1 SHARP (as indicated by the reception control), the signal is fed through the sharp filter. This filter has a bandpass of 3.2 kc centered on the 200-kc intermediate frequency. When S2A is set at A2 or A3 SHARP the medium filter is selected. The bandpass of the medium filter is 8 kc centered on the 200-kc intermediate frequency.

The output of the selected filter is amplified in the three i-f amplifiers. The first two stages are controlled by a section of the r-f gain control. The output of the third i-f amplifier is fed to the diode detector, the BFO mixer, and a cathode follower.

The diode detector receives the i-f signal during m-c-w reception, feeding its output through switch S2D to the noise limiter. During normal signals the output of the noise limiter is developed across the output control. When a noise pulse of greater than normal amplitude arrives, the noise limiter automatically cuts off the audio output for the duration of the noise pulse. During c-w and frequency-shift reception, the signal received at the BFO mixer is heterodyned to produce a signal in the audio range. Its output is passed through S2B to the output control.

Tuning meter M1 in the detector circuits shows the diode detector output current and gives a visual indication of tuning.

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The cathode follower of the second i-f assembly is provided to supply frequency-shift signals to a frequency-shift converter.

For c-w and m-c-w reception (position A1 and A2) the input to the first audio amplifier is obtained directly from S2E. In these positions of S2 the output control is inactive. Frequency-shift signals are applied to the first audio amplifier from the arm on the output control.

The second audio amplifier receives the output of the first audio amplifier either directly or by way of a narrow bandpass filter. During A1 reception S2F is placed in the A1 SHARP position to insert the filter and decrease adjacent channel interference. The A1 BROAD, A2, and f-s-k positions of the switch bypass the filter.

The output of the second audio amplifier is applied to two series limiters. During A1 and A2 reception, alternate limiters clip the signal equally on the positive and negative half cycles. When the reception control is switched to the f-s-k position, the series limiters are inactive.

The signal is then fed to a driver amplifier, which increases the signal strength required to drive the beam power output amplifier. The latter stage feeds the output transformer, T9. The secondary of the transformer supplies a balanced 600-ohm line through its terminals, 5 and 7, and applies the signal to the phone jack through an unbalanced winding connected between terminals 3 and 4. The signal developed in the unbalanced line is also applied to the output meter, M2, which indicates the signal level in decibels. Multipliers are switched in series with the meter to increase the decibel range that may be covered. Zero db corresponds to a power level of 6 mw through a 600-ohm load.

The power supply used in the AN/SRR-11 is a conventional full-wave supply, and will be considered in the discussion.

The crystal calibrator circuit provides accurate frequency check points, which may be used for calibrating the main tuning dial. The check points occur at every 10 kc throughout the tuning range of the AN/SRR-11. A cal switch, S5, is connected so that, in the OFF position, the B voltage is removed from the calibrator. When placed in the ON position, S5 allows the B voltage to be applied to the calibrator, and removed from the preamplifier and the BFO. This prevents interference with the crystal calibrator from either externally applied signals or the BFO.

R-F AMPLIFIER SECTION

A simplified schematic of the r-f amplifier is illustrated in figure 4-4.

Band Selector

The band selector, S1, permits the selection of any one of five bands. The signal circuits are similar in all five bands except for the magnitudes of the inductance and capacitance. To keep circuit analysis simple, only one band is shown in the figure. The four sections of S1 are ganged together and controlled by the band selector knob located on the receiver front panel. The center shaft of the switch (shown on the schematic) is electrically insulated from, but mechanically connected to, its associated wiper arms. The three wiper arms make sliding contact with their respective segments.

The components chosen by the band selector determine the frequencies that will be passed into the receiver from the antenna. On band 1 of the AN/SRR-11, the lowest frequency range is covered as shown in table 4-3. The band selector therefore connects the inductance and capacitance grouping, which produces the largest L-C product. On band 5, the highest frequencies are to be received, and the band selector connects the components with the smallest L-C product. Continuous tuning within the band is accomplished through the use of the ganged sections of the tuning capacitor, C5.

Both the input and output transformers in the r-f amplifier section of the receiver are changed for each respective band. The tuning capacitor, C5, is not changed by the band selector, and therefore provides tuning on all bands.

Antenna

The antenna used with the AN/SRR-11 aboard ship is usually of the long-wire or whip type. The capacity of the antenna used should be within the range of 30 to 5000 $\mu\mu$f. Capacitor C1 (fig. 4-4) may be connected in series with the antenna input line when an antenna of high capacity is used to reduce the total capacity. Microswitch S7, operated by a cam on the shaft of the Antenna Compensation Control, is closed for 180 degrees of rotation of the control. In this position, designated low, C1 is out of the circuit, and the primary circuit capacity corresponds to that of the antenna. The switch, S7, is open for the second 180 degrees of rotation

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of the Ant Comp control, and C1 is switched into the antenna input circuit. In this case (designated high) the total input capacity is that of the series capacitances formed by C1 and the capacity of the antenna used.

The antenna section of the ganged tuning capacitor, C5A, is shunted across the secondary of the antenna transformer through switch S1A. A section of the switch, S1 (not shown) shorts the unused transformers to prevent interference with the band chosen. Four of the transformers will be shorted at all times, while the other serves to tune the receiver to the desired signal. The method of shorting the unused transformers is similar to that described for the power amplifier circuit in chapter 6 of this training course.

Two links, 01 and 02, make possible the matching of the antenna transformer to a high or low impedance input to the r-f amplifier. The two links are mechanically ganged together. In the high Z position (position 1) the antenna is connected through S1A to a tap on the secondary of the chosen transformer. The input voltage is developed from the center tap of the transformer through R3 to ground. R2 serves as an antenna input divider resistor, which attenuates all input frequencies, particularly the weaker signals on either side of the frequency to which the receiver is tuned. R3 acts as the grid return. The autotransformer action of the secondary of T1 develops the input voltage, which is applied through a section of the bandswitch, S1A, to the control grid of V1.

Antenna Preamplifier

The chosen frequencies are amplified in V1 and passed through a section of the band selector, S1B, to the primary of the plate transformer, T2. Note that the plate potential is lower than that of the screen by the amount of the d-c voltage developed across the primary of transformer, T2. The difference in the two potentials is small since the d-c resistance of the T2 primary is very low. At any rate, with the screen at the higher potential, the current passing through the tube is increased. This is true because the screen, being located closer to the cathode than the plate, has the greater effect on the acceleration of current through the tube. The variations in plate current are therefore greater than that of a conventional r-f amplifier, resulting in a greater output signal amplitude. The primary of T2 is shunted by C8 in order to equalize the receiver gain over the band.

The r-f gain control is divided into two sections, one section of which controls the r-f amplifiers, while the other controls the i-f amplifiers. L1, in series with the r-f gain control, serves as a filter, which prevents r-f variations from entering the biasing circuit. C28 bypasses both the coil and resistor, and therefore assists in keeping the voltage across the gain control constant. The cathode of the r-f amplifier, V2, is tied to the junction of L1 and R6. Through this connection, the gain control potentiometer, R20A, provides variable bias for controlling the gain of V1 and V2.

R-F Amplifier

The output of the antenna preamplifier is coupled across the selected transformer to the control grid of the r-f amplifier, V2. The signal is applied through a section of the band selector directly to the control grid. The absence of a coupling capacitor and resistor at the input of V2, as may be found in some amplifiers, prevents the loss of some portion of the input signal across these components. For this reason, the omission of grid components is quite common in r-f and preamplifier stages.

A section of the ganged tuning capacitor, C5B, is shunted across the secondary of the selected transformer, T2, by a section of S1B. Through the use of this capacitor, continuous tuning is provided over the selected frequency band. The secondary coil of T2 is tuned by a variable slug core. Capacitor C11 serves as a trimmer for the V2 input circuit. The signal voltage is taken from the secondary center tap and applied through a section of S1B to the control grid of V2. A section of the band switch (not shown in the schematic) is incorporated in the receiver to short all of the unused transformers.

As was stated earlier, the r-f gain control is connected so that it also affects the bias of V2. Most military receivers are equipped with r-f gain controls, because their places of operation are so widely varied. When the receiver is used close to or in a port, the gain control may be set to place maximum bias on the r-f stages, thereby decreasing their gain. On the other hand, when the receiver is used far out at sea, the bias may be decreased to a minimum, to allow the maximum amplification in each of the r-f stages.

When the gain control is set so as to reduce the gain an appreciable amount, signals may be received, which are strong enough to cut off V2.

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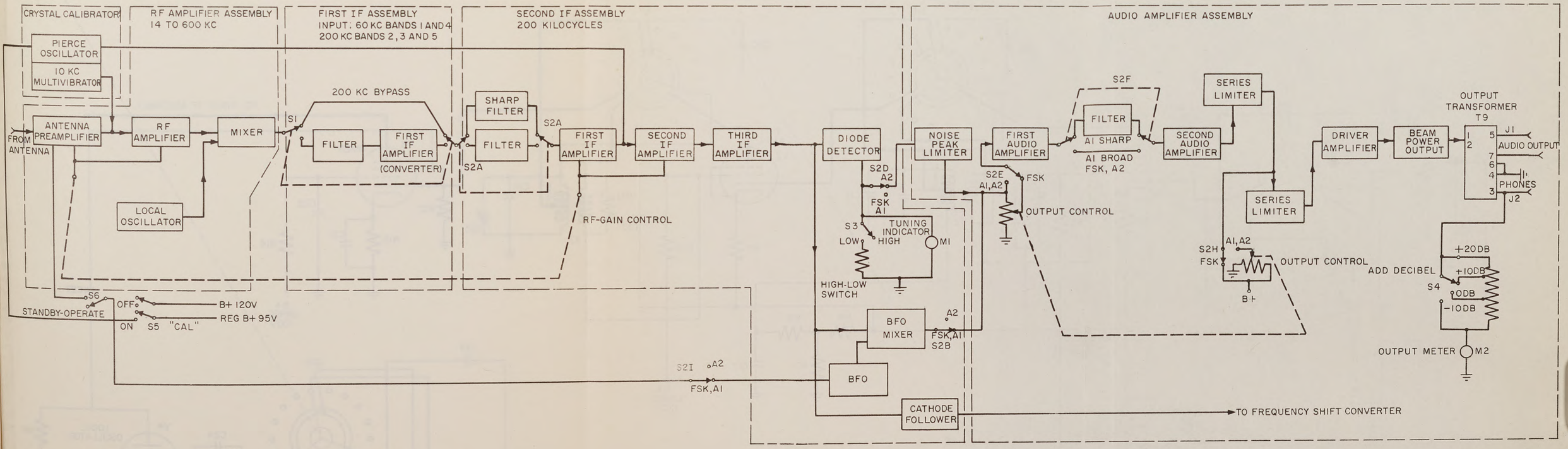

Figure 4-3. —Block diagram of AN/SRR-11.

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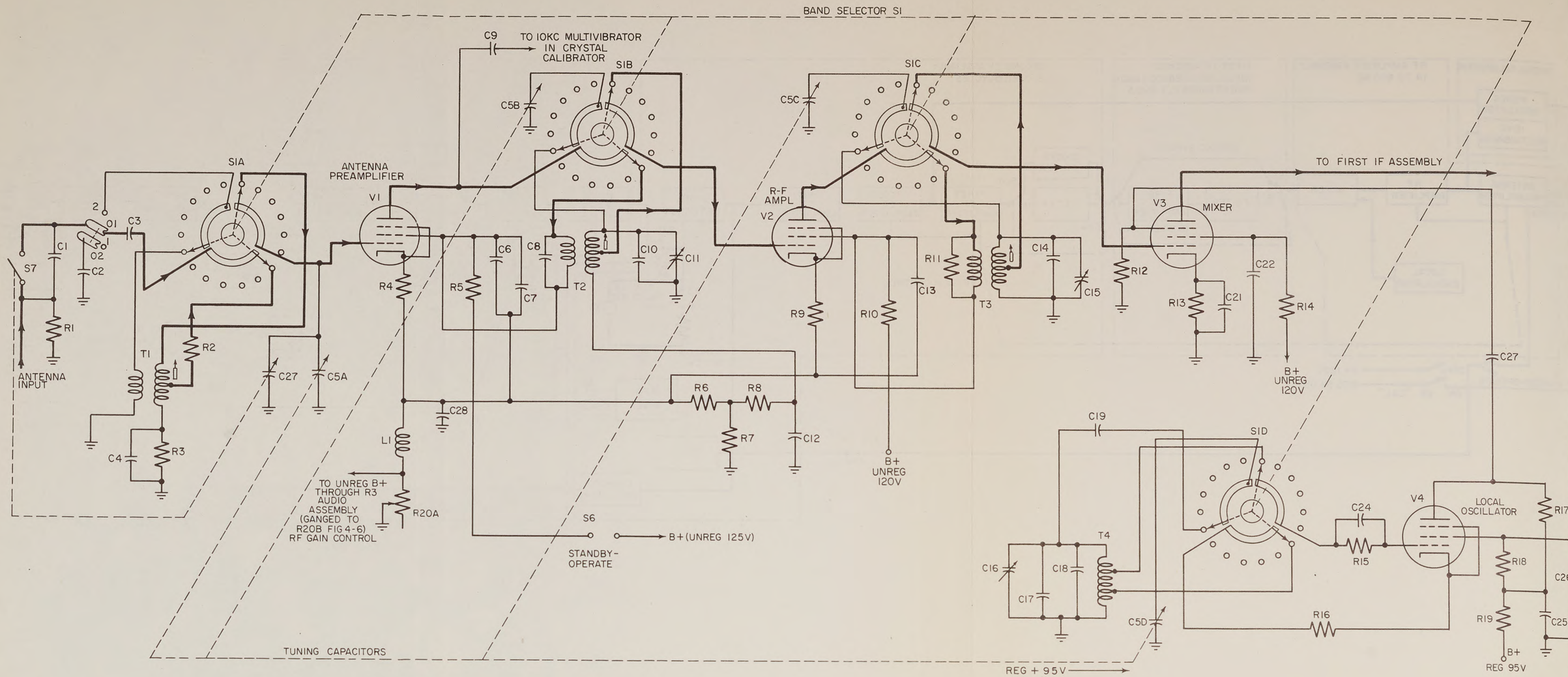

Figure 4-4. —Radio frequency amplifiers, simplified schematic.

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r this reason, a slight positive potential across
7 is applied through R8 and made available at
e grid of V2 through a section of the band
lector. Thus, the possibility of the tube being
iven to cutoff by strong r-f signal is reduced.
and C12 form a grid decoupling network,
ch prevents r-f variations from entering the
sing circuit from the grid of V2.

xer

The input to the mixer, V3, control grid is
ained from the secondary tap on the associ-
d transformer, T3, as chosen by the band
lector. Section C5C of the ganged tuning cap-
tor serves to tune the T3 secondary to the
nal frequency. C15 is the trimmer capacitor
ross the T3 secondary tuned circuit to fine tune
secondary.

cal Oscillator

The oscillator, V4, used to produce the heter-
yning frequency is an electron-coupled Hartley
cillator whose cathode and control grid are
ced in the oscillator circuit by the S1D sec-
n of the band selector. In this manner, the
cillator frequency changes in accordance
th the chosen grid-cathode components. T4,
oscillator transformer, is shunted by fixed
pacitors C17 and C18 and trimmer capaci-
C16. C5D is the main tuning capacitor. Grid
k bias is supplied to the oscillator by the
ion of R15 and C24.

Feedback sufficient to sustain oscillations
provided through capacitor C26 and developed
the grid of V4 by the autotransformer action
T4. The oscillator signals are electron
pled to the plate from which they pass
ough capacitor C27 to the suppressor grid
the mixer, V3. The two frequencies are
erodyned in the mixer stage to produce the
rmediate frequency. The oscillator tracks
the high side of the signal frequency, so
t it is 60 kc higher than the signal fre-
ncy on bands 1 and 4, and 200 kc above the
nal frequency on bands 2, 3, and 5.

FIRST I-F ASSEMBLY

ter

The switches, S1E and S1F (fig. 4-5) are
ged to the band selector switch so that their
sitions correspond to the band chosen in the
f section. These switches, like those already discussed, are shown in the BAND 1 position. They are responsible for the bypassing of the first i-f assembly in bands 2, 3, and 5; for feeding the input to the filter, Z1, when the receiver is tuned to bands 1 or 4 (position shown); and for applying operating potentials for the mixer and the first i-f assembly, as determined by the setting of the band selector.

On band 1, when the output of the mixer is 60 kc, the first i-f assembly is not bypassed, since the 60 kc must be heterodyned with a 140-kc frequency in order to produce the desired intermediate frequency of 200 kc. The mixer, V3 (in the r-f section of figure 4-4) receives its plate voltage through the filter, Z1, and contacts 1 and 6 of the A section of S1E. The mixer input is fed through contact 6 of S1E, and through its shorting ring to contact 1. From this point, the signal is fed to the 60-kc filter, Z1.

Converter

The oscillator used to produce the 140-kc heterodyning frequency is connected as an electron-coupled Pierce circuit, since its crystal, Y1, is connected between the screen (which serves as the plate of the oscillator) and the control grid. The oscillator, V5, receives its plate (pin 5) potential through contacts 6 and 1 of S1F when the band selector is placed in the BAND 1 position. Likewise, on band 1, screen voltage is supplied through contacts 7 and 1 of S1F. R21, bypassed by capacitor C30, develops the grid leak bias, while R22 serves as a safety bias resistor, which protects the tube, V5, in case oscillations fail. Y1 is a temperature-compensated quartz crystal, which produces the 140-kc frequency.

If the band selector switch is set so that the receiver operates on band 4, S1F will have advanced 3 steps, and the plate potential will now be applied through its contacts, 6 and 4. The screen potential is applied, in this case, through contacts 7 and 4 of S1F. You should note that S1E now couples the input from the mixer, V3 (in the r-f assembly) through contacts 6 and 4 to the 60-kc filter, Z1.

Since the first i-f assembly is bypassed on bands 2, 3, and 5, the bypassing connections through the switches, S1E and S1F, are discussed relative to the BAND 2 position only. The input from the mixer in the r-f assembly (when band 2 is selected) bypassed V5 through contacts 6 and 2 of the A section of S1E, to contacts 2 of the A section of S1F, through its

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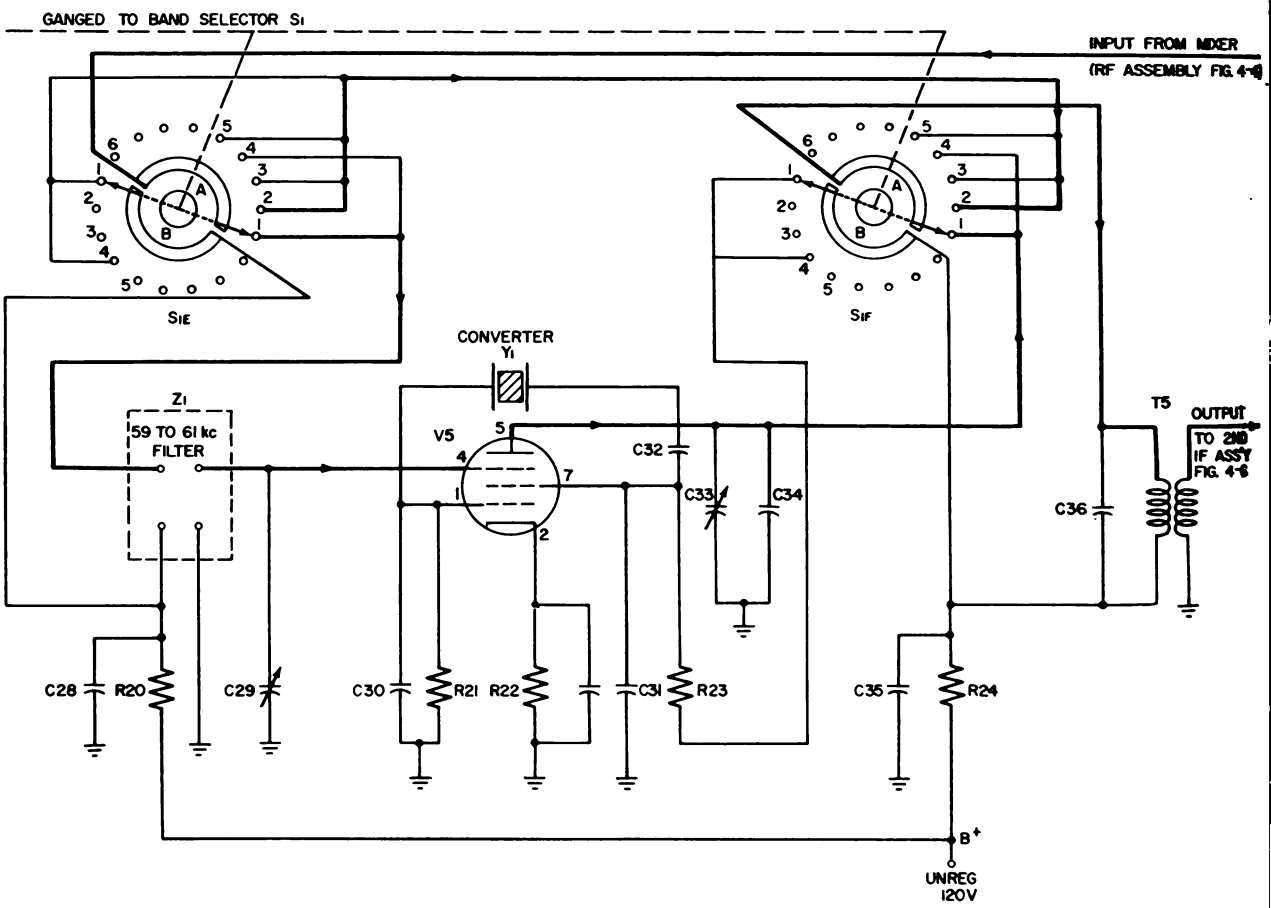

Figure 4-5. —First I-F Assembly AN/SRR-11, simplified schematic.

shorting ring to contact 6, and thence to the tuned primary circuit of the output transformer, T5. The filter and oscillator circuits are therefore bypassed. It should be noted that in the BAND 2 position of the band selector switch, operating potentials are removed from the plate and screen grid of the oscillator tube, V5, by switch section S1F. Also note that the plate potential is now applied to the mixer (in the r-f assembly) through contacts 6 and 2 of the A section of S1F and contacts 2 and 6 of the A section of S1E.

The 60-kc filter, Z1, allows the passage of a band of frequencies from 59 to 61 kc. The output, fed to the V5 suppressor grid, is heterodyned with the 140-kc signal, which is generated in the V5 oscillator section. Thus on bands 1 and 4 the desired 200-kc intermediate frequency is supplied to the output transformer, T5, as the sum of 60 kc and 140 kc.

SECOND I-F ASSEMBLY

Filters

Reference to the block diagram of figure 4-3 shows that the second i-f assembly includes three stages of 200-kc i-f amplification, a BFO and BFO mixer, a diode detector, and a cathode follower. Also included in the second i-f assembly are two filters, Z2 and Z3, designated sharp and medium respectively. The filter to which the output of the first i-f assembly is applied is determined by the setting of the switch, S2A. All of the switch sections, S2A, S2B, S2C, and S2D, are ganged to the reception control on the front panel and determine the mode of operation of the receiver.

The output from the first i-f assembly passes through the A section of switch S2 via contact 1 for FSK, (position shown) contact 2 for A1 broad, and contact 3 for the A1 sharp selection. The signal then passes through the

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"sharp" filter, Z2, which allows the passage of frequencies ±1.6 kc centered around the 200-kc intermediate frequency. When the reception control is set at A2 BROAD, (position 4) the "medium" filter, Z3, is selected. Z3 allows passage of frequencies within ±4 kc of the intermediate frequency. The S2C and S2D sections of the reception control will be considered when their operation is pertinent to the immediate discussion.

Amplifiers

Signals received at the control grid of the first i-f stage, V6 (figure 4-6) are amplified in accordance with the setting of the r-f gain control, R20B. The A and B sections of the r-f gain control are ganged together and controlled by the gain control on the receiver front panel. The A section, R20A, as forestated, controls the gain of the antenna preamplifier and r-f amplifier shown in figure 4-4. The B section, R20B, is connected in series with the cathodes of the first and second i-f amplifiers of the second i-f assembly so that it controls the gain of these two stages.

The three i-f transformers, T6, T7, and T8, are overcoupled to produce broadly tuned (coupled) circuits between the stages. The extension, P, of the primary coil, is placed in close proximity to the secondary coil, S, thereby increasing the coupling.

R51, in parallel with the V8 cathode biasing network, is a temperature compensating resistor. When the equipment is initially placed in operation, R51 acts as a low resistance across R39 reducing the bias on V8, making the set come into normal operation faster. As the resistor becomes heated, its resistance value rises, thereby increasing the bias and decreasing the gain of the amplifier.

The output of the third i-f amplifier feeds four circuits: the diode detector, V9; the tuning meter, M1; the BFO mixer circuit, V11; and a cathode follower, V12.

Diode Detector

During A2 reception the 200-kc i-f signals are applied by way of T8 to the diode detector, V9. The diode load resistor, R43, develops the audio output. The purpose of links, 01 and 02, is discussed later. When the switch, S2C, is rotated so that a circuit is completed through contacts 4 and 5 of section C (as will be the case during A2 reception), the diode detector feeds its output to the noise limiter in the audio amplifier assembly.

Tuning Meter

The tuning meter, M1, in the detector cathode indicates the signal strength at the detector output. When the desired signal is detuned, only a small rectified current passes through V9, and the meter reading is low. The reading increases as the desired signal is properly tuned in. The high-low switch, S3, in the LOW position attenuates the meter reading by connecting a shunt resistor, R42, across the meter. S3 is a spring-loaded switch, which is normally in the LOW position. When the switch is moved to the HIGH position and released, the spring acts to return it to the LOW position. This is done to protect the meter.

Beat Frequency Oscillator

The beat frequency oscillator comprising V10 provides the heterodyning signal to the BFO mixer, V11, when the reception control is placed in the A1 BROAD, A1 SHARP, or FSK position. The heterodyning process then produces a frequency in the audio range.

V10 is an electron-coupled Hartley oscillator. Its output frequency is adjusted by capacitor C42 so that after heterodyning in the BFO mixer, the output frequency is variable in the range of ± 1500 cps on either side of the zero beat frequency. The frequency vernier on the receiver front panel controls the capacitance of C42, and is therefore used to adjust the frequency of the output audio note. C41 serves as a trimmer capacitor. Operating voltage for V10 is received through contact 1, 2, or 3 or S2D.

BFO Mixer

The output of the oscillator is coupled to the suppressor grid of the BFO mixer, V11, where it is heterodyned with the signal from the last i-f stage. The latter signal is coupled to the control grid of V11 by C64. During A1 BROAD, A1 SHARP, and FSK reception the two signals are heterodyned in V11. After heterodyning, the resultant difference frequency is fed to the audio amplifier assembly through the A section of S2C of the reception control.

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Cathode Follower

The 200-kc intermediate frequency may be fed to a frequency-shift converter, which requires the i-f signal input for operation. The i-f output is coupled through C69 to the control grid of the cathode follower, V12. The signal is developed across the cathode resistor, R52, and coupled by C70 to a coaxial receptacle on the back of the receiver. R63, in the plate circuit of V12, is bypassed by C71 to form a decoupling network.

Present day frequency-shift converters such as the AN/URA-8B (discussed in chapter 8 of this training course) require audio frequency input signals. In this case, the output to the converter is taken from the receiver output transformer through a balanced audio line of 600 ohms. The mark and space frequencies are correctly set by adjusting the beat frequency oscillator after the receiver has been tuned to the desired carrier frequency.

AUDIO AMPLIFIER ASSEMBLY

The audio assembly (fig. 4-7) is identical for the three receiver types. Certain connections in the audio assembly are slightly modified for use in the AN/SRR-11 receiver. These differences are discussed in this section.

Input Circuit

Input to the audio assembly is determined by the setting of S2E, which is ganged to, and controlled by, the reception control. When the latter control is set to the FSK, A1 BROAD or A1 SHARP position, the output from the BFO mixer is applied to the control grid of V13 through contacts 2, 3, or 4 of the A section of S2E. The input signal is developed across R56B. R56A and R56B are ganged together and controlled by the OUTPUT control on the receiver front panel.

Noise Limiter

V15 is a series noise limiter for A2 reception. The V15 current is caused to vary in accordance with the audio input via the detector input, A2. The output developed by R64 at the V15 cathode (V15 normally conducting) is coupled through R60, C73, and contacts 4 and 5 of S2E to the V13 grid. When the reception control of the AN/SRR-11 is set at A2 (or the A3 position on the AN/SRR-12 and 13), the input from the detector (V9 of figure 4-6) is developed across R62 and R63. The latter two resistors are paralleled by a filter comprising R61 and C74. The filtered input signal is used to provide the operating potential for the cathode of the noise limiter, V15. The plate voltage is developed from the junction of R62 and R63 to ground, while the cathode potential is the same as that appearing across C74.

Since the input voltage, after being filtered, is effectively d-c, the value of the voltage developed across C74 is almost equal to the peak value of the signal input. The plate voltage usually less negative than the cathode is derived from a purely resistive path, and may take on sudden or instantaneous changes. These changes may appear as noise pulses in the input. The cathode voltage, which is developed across C74 cannot respond to sudden voltage changes due to the charge time of the capacitor. Under normal operating conditions, V15 conducts a current that varies in accordance with the A2 signal across R63.

When a noise pulse (or modulated signal that exceeds normal modulation percentage) is detected, the V15 plate voltage goes more negative. The V15 cathode voltage, however, remains almost constant because of the filtering action of C74 and R61. Thus, for a short time interval the plate is more negative than the cathode, and during this time the output across the V13 grid is discontinued. In this manner, the output noise pulse is limited to a value corresponding to the maximum normal modulation percentages determined by the relative values of R62 and R63. In the AN/SRR-11, 12, and 13, the noise limiter is set to limit impulses to a level corresponding to 40% modulation.

When S2E is set to the FSK position (as shown), R56B provides a means of adjusting the signal level before it is applied to the V13 grid. In the other positions of the switch (A1 BROAD, A1 SHARP, and A2) the B section of the output control is inoperative since the signal output is taken across the entire resistor. The full value of R56B acts as the grid resistor for the V13 grid in all modes of operation except FSK. For FSK signals the V13 grid is connected between the arm of R56B and ground.

First Audio Amplifier

The first audio amplifier normally has a broad band characteristic. However, a narrow range of audio frequencies are obtained in the

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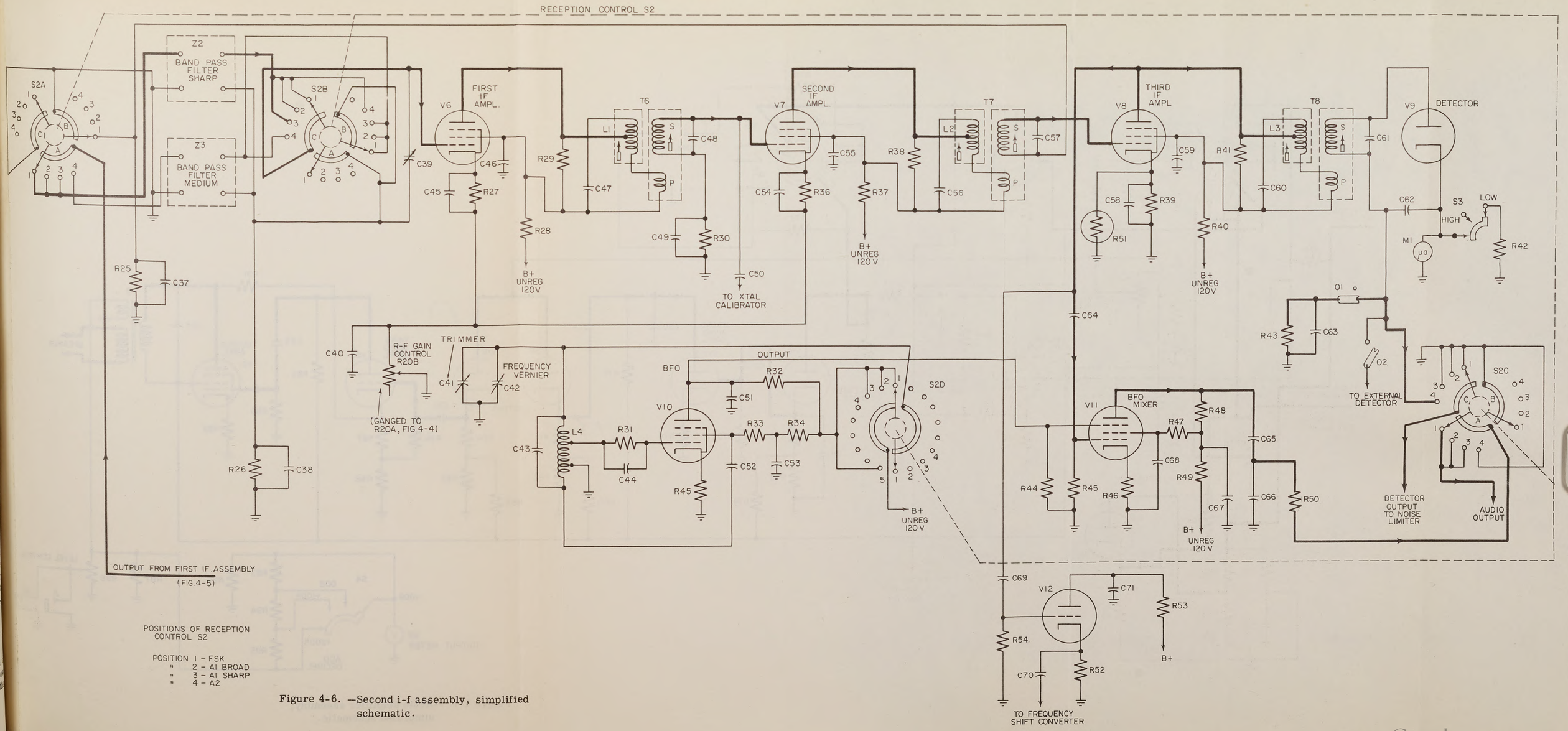

Figure 4-6. —Second i-f assembly, simplified schematic.

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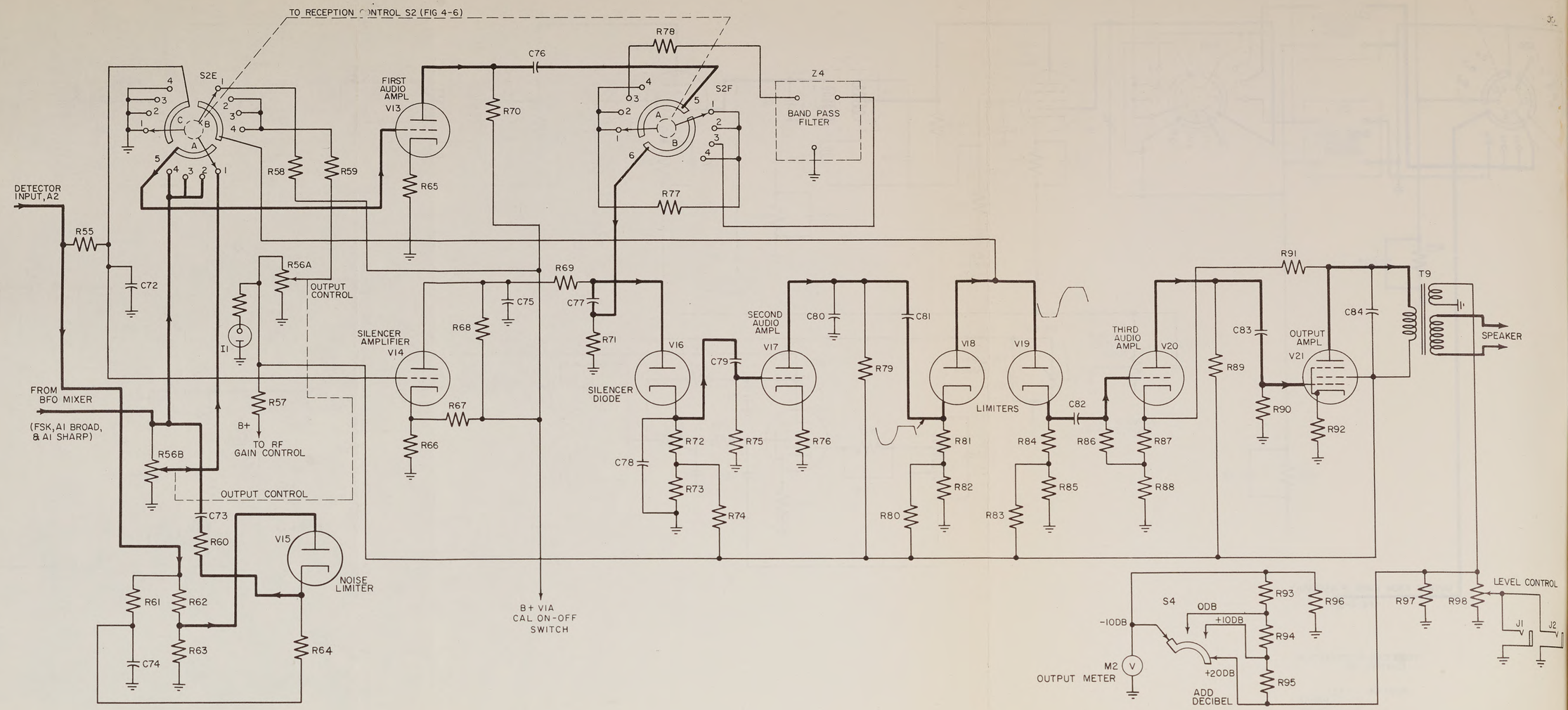

Figure 4-7. —Audio amplifier assembly, simplified schematic.

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V13 output circuit when the reception control is set to the A1 SHARP position. The output of V13, in this case, is coupled by way of C76 through contacts 5 and 3 of section A of S2F to R78 (an attenuating resistor) and to a bandpass filter, Z4. Z4 then passes frequencies ranging between 240 and 4300 cps with maximum response at 1000 cps. The output of Z4 is passed through contacts 3 and 6 of the B section of S2F to the silencer circuit.

Second Audio Amplifier

The output of the first audio amplifier is coupled to a silencer circuit comprising tubes V14 and V16. The silencer diode, V16, is connected in series between the first and second audio amplifiers V13 and V17. Since the silencer circuit is used only in the AN/SRR-12 and 13 receivers, its connections in the AN/SRR-11 equipment are such that V16 is always conducting. In this manner, the audio output of V13 (and the filter Z4 when used) is coupled through C77, the conducting diode V16, and through C79 to the second audio amplifier, V17. A detailed explanation of the silencer circuit is given at the end of this chapter.

V17 is a conventional audio amplifier. Degeneration is permitted at the cathode by the unbypassed cathode resistor, R76, to improve the fidelity of the audio output.

Diode Limiters

Two series diode limiters, V18 and V19, are provided to clip equally on the positive and negative portions of the audio cycle. When switch S2E of the reception control is set for the A1 or A2 positions (corresponding to the positions 2, 3, or 4) the plate potential of the limiters is obtained from the A section of the output control, which is connected to the B supply. R59 is a series dropping resistor for the two diodes.

With the plate potential of the two diodes set at a certain positive potential, V18 (which receives its input at its cathode) is allowed to pass the full negative swing of the audio signal. However, during the positive half cycle of the input, the positive cathode potential (with respect to ground) may exceed that of the plate. In such cases, V18 is cut off during a portion of the positive half cycle, thereby limiting or clipping a part of the positive input signal. Likewise, during the negative half cycle, the value of the negative input may be sufficiently strong to overcome the plate to ground potential of V19. In this way, the output signal, as it appears at the cathode of V19, is clipped on the negative half cycle. The level at which the signal is clipped is determined by the setting of the A section of the output control, R56A. The control may be set so that the diodes provide an output-input voltage gain up to 100, or over a range of 40 db.

When the reception control is set to the FSK position (position 1 of S2E or the A3 positions of the AN/SRR-12 and 13), the plate potential of the diodes is fixed by a connection to the B supply through R58. The value of the plate voltage in this case is such that the diodes remain conductive.

Audio Output Stages

The signal voltage is taken from the cathode of the second limiter, V19, and applied through C82 to the control grid of the audio driver stage, V20. The grid resistor of V20 is connected to the junction of R87 and R88, so that the positive potential (with respect to ground) at this point will allow a larger negative input signal amplitude before driving V20 to cutoff.

The output of the beam power pentode, V21, feeds the primary of the output transformer. A negative (degenerative) feedback voltage is fed from the plate of V21 to the cathode of V20 by the voltage divider comprising R91, R87, and R88. The amount of this voltage is determined by the gain of V21.

The feedback voltage at any instant is such that it opposes the signal appearing at the control grid of V20. As the secondary of the output transformer is loaded, the output of the power amplifier decreases. The secondary load may be a set of headphones at J1. If additional headphones are inserted at J2, the secondary of T9 experiences a load increase, and the output of V21 decreases. With the decrease in the output of V21, less degenerative voltage is fed back to V20, and a larger input will be applied at the control grid of V21. From this we conclude that the feedback voltage is provided to maintain a nearly constant output under changing load conditions.

Output Circuits

The signal to the output meter, M2, may be decreased in four steps by the add decibel switch, S4. This permits the meter to indicate, in decibels, a wide range of output levels. Zero db of the meter circuit corresponds to a power

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level of six mw into a 600-ohm load. Multiplier resistors R93, R94, and R95, are selected by S4 for the +20-db, +10-db, 0-db, and -10-db positions. No attenuation is provided when the switch is in the -10-db position (as shown) so that the low output levels can be read. The -10-db position of the add decibel switch is spring loaded, and the switch will fall back to the 0-db position if the switch arm is not held in the -10-db position. R96 serves as a shunt for the output meter. R97 and R98 in parallel develop the voltage for the headphones and meter circuits. R98 provides a method of adjusting the audio level of the headphones.

CRYSTAL CALIBRATOR AN/SRR-11

The crystal calibrator circuit (fig. 4-8) provides accurate frequency check points for calibrating the main tuning dial of the receiver. These check points occur at every 10 kc throughout the tuning range of the receiver. The 10-kc crystal calibrator contains a 50-kc Pierce oscillator comprising V22 and a 10-kc multivibrator, V23.

50-KC Pierce Oscillator

The crystal, Y2, and capacitor C87 are the main frequency determining components of the oscillator. The components maintain the desired oscillator frequency of 50 kc. In addition, C87 protects the crystal from the d-c plate potential of V22.

The feedback voltage for the oscillator is passed through C90 and developed at the control grid of V22 by C88 and C89. C88 is made adjustable so that the frequency of the oscillator may be corrected (as required) over a limited range. This capacitor should not be mistaken for a tuning capacitor, since a Pierce oscillator does not require tuning. Furthermore, C88 is properly adjusted at the factory and should not require field adjustments. C89 works in conjunction with R102 to provide the necessary grid leak bias for the oscillator.

The operating potentials for the tubes of the crystal calibrator are obtained from a regulated B supply. This is done to improve the oscillator stability, which is necessary in any circuit used as a frequency check standard.

The oscillator provides two output signals. One of the outputs is taken from the cathode of V22, and is coupled through C91 to the control grid of the 10-kc multivibrator, V23. Another output (50 kc) is taken from the plate of V22 and coupled through C93 to the second i-f assembly (fig. 4-6). In this unit, the signal

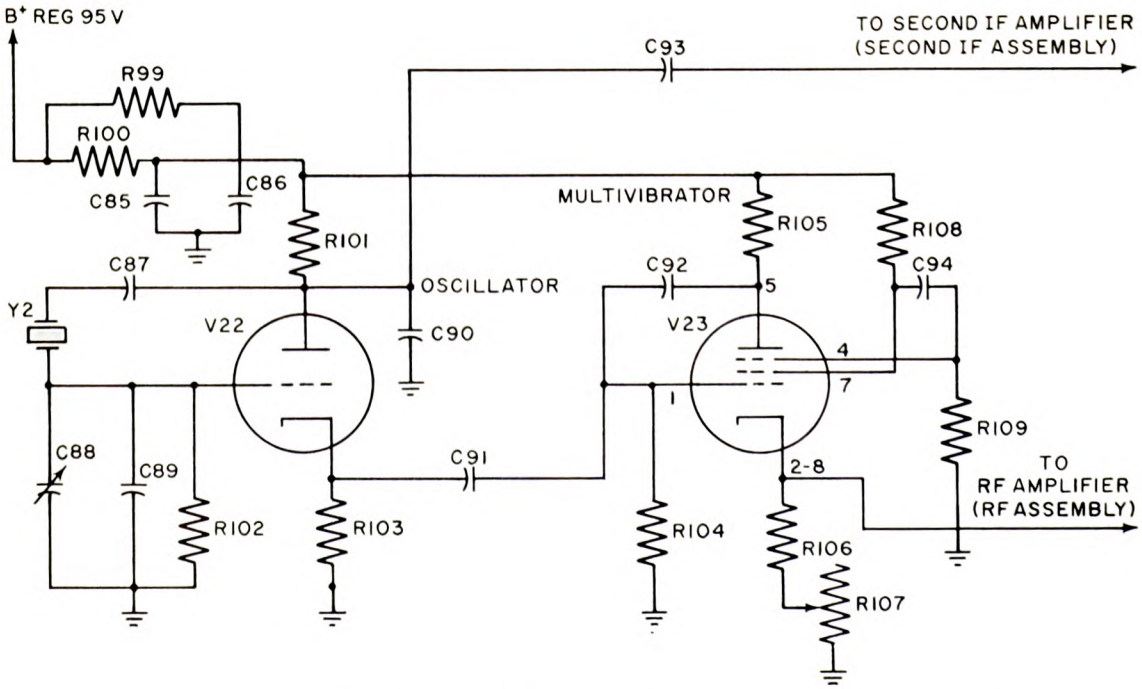

Figure 4-8. —Crystal Calibrator AN/SRR-11, simplified schematic.

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passes through C50 to the control grid of the second i-f amplifier, V7.

10-KC Multivibrator

The 10-kc multivibrator, V23 (fig. 4-8) is synchronized by the 50-kc signal from the cathode of V22. Pentode V23 is connected as two triodes to form a plate-coupled multivibrator. The cathode (pins 2 and 8) the control grid (pin 1) and screen grid (pin 7) act as the first triode. The control grid performs its normal function, while the screen grid serves as the plate of this section of the multivibrator. The common cathode, the suppressor grid (pin 4), and the plate (pin 5) make up the second section of the multivibrator. In the latter triode section, pin 4 acts as the control grid. The second triode may be considered to have a virtual cathode existing between pins 7 and 4 since this is the area from which its electrons are obtained.

The second triode plate (pin 5) is coupled to the control grid of the first triode (pin 1) through C92 and R104. The first triode plate (pin 7) is coupled to the second triode control grid (pin 4) through C94 and R109.

Assume that the second triode is conducting. If a positive half cycle of the 50-kc input from V22 arrives at the control grid (pin 1) of V23 at this time, the first triode section begins to conduct. Its screen potential drops, causing a decrease in the voltage across C94. This voltage change is developed across R109 such that it decreases the current through the second section and, likewise, the self-bias developed across R106 and R107. The plate potential of the second triode increases, and charges C92 through R104. The positive potential (developed from control grid to ground) causes the first triode to conduct an additional current. This also causes a further drop in the screen potential of the first triode, and an increase in plate potential of the second. The process is continuous and rapidly accumulative until the first section conducts saturation current, and the second section of V23 is cut off.

The arrival of a negative half cycle of the 50-kc input from V22 at this time (the first triode conducting and the second cut off) will reduce the current through the first triode section. This will be evidenced by a rise in the screen (pin 7) potential. The instantaneous screen voltage rise is sufficient to begin the charge of C94 through R109. This action tends to make the grid of the second triode section positive with respect to ground. The resultant drop in plate potential of the second triode allows C92 to discharge so that it aids the negative-going signal being received at the control grid. The first section therefore moves further toward cutoff while the second section moves to saturation.

The circuit constants of the multivibrator are such that the multivibrator is synchronized to 10 kc by the 50-kc signal from the Pierce oscillator. The cathode resistance (which determines the bias of both triode sections) may be adjusted by R107 to assure the desired 10-kc output frequency.

The 10-kc multivibrator output is coupled from the cathode of V23 to the r-f assembly (fig. 4-4) via C9 to the plate of the antenna preamplifier, V1. The input square wave signals are rich in harmonics. One of these harmonics will be heterodyned in the mixer stage, V3, with the signal from the local oscillator, V4. The heterodyning signal, as it is supplied to the mixer, is determined by the settings of the band selector, S1, and the main tuning capacitor, C5.

The fourth harmonic (200 kc) of the 50-kc output (fig. 4-8) is taken directly from the oscillator plate, and applied through C93 (located in the crystal calibrator) to the second i-f assembly (fig. 4-6). C50 then couples this signal to the control grid of the second i-f amplifier, V7. The two signals (one passing through the first i-f amplifier of the second i-f assembly and the other from the crystal calibrator) will give a zero beat at every 10-kc separation of the tuning dial setting.

If the cal switch on the receiver front panel (fig. 4-2) is turned to the ON position, the main tuning dial can be calibrated independently of external signals by the use of the cal adjust control on the receiver front panel. The cal adjust control provides a means of shifting the frequency scale on the translucent screen when recalibrating the receiver in conjunction with the crystal-controlled calibrator. This is done by a slight, horizontal movement of the lens system.

POWER SUPPLY AND POWER SWITCHING CONTROLS

Power Supply Circuit

The power supply (fig. 4-9) is identical for the three receivers. A dial indicator lamp (not shown) is supplied by the a-c voltage between terminals 1 and 3 (ground) of transformer T10.

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Ballast tube R110 is used to regulate the filament voltage of the local oscillator (fig. 4-4) in order to improve frequency stability. The voltage drop across this resistance may vary from 4.5 to 11.2 volts. Hum bucking resistor R111, is shunted across the filaments of the noise limiter, V15 (fig. 4-7). The resistor functions to balance out a-c hum, which might otherwise be coupled through the filament-to-cathode capacitance of the noise limiter. The composite audio signal from the noise limiter containing the hum would then be introduced to the control grid of the audio amplifier and amplified in the plate circuit. The electrical center of the filament transformer is +120 volts with respect to ground. When the tap of R111 is at its center position, the voltage change on each side of the filament is made equal and opposite, and the hum is reduced.

A regulated +95 volts is obtained from the plate of the voltage regulator tube, V26 (fig. 4-9). This voltage is supplied to the crystal calibrator, the local oscillator in the r-f assembly, and the AGC circuit. The AGC circuit, however, is not used in the AN/SRR-11. The remaining circuits of the receiver are supplied by the unregulated +120 volts obtained from the junction or R113 and R114.

Calibrator and Standby-Operate Switches

The cal switch, S5 (fig. 4-10) is connected so that the OFF position (as shown) the B supply voltage is not applied to the crystal calibrator. With the switch in this position, the receiver supplies its normal output in accordance with the setting of the band selector and reception control, as previously discussed. In the ON position of S5, the regulated B supply voltage is applied to the calibrator, and unregulated B+ is removed from the BFO (V10 of the second i-f assembly in figure 4-6) and from the antenna preamplifier (V1 of the r-f assembly in figure 4-4). The B supply voltage is removed from these stages to prevent interference in the receiver (from external or internal sources) during the time frequency calibration of the main tuning dial is taking place.

The antenna preamplifier receives its operating voltage through the contacts of the standby-operate switch, S6 (fig. 4-10) in the ON position. When S6 is open, the antenna preamplifier is inoperative, and the receiver output is discontinued.

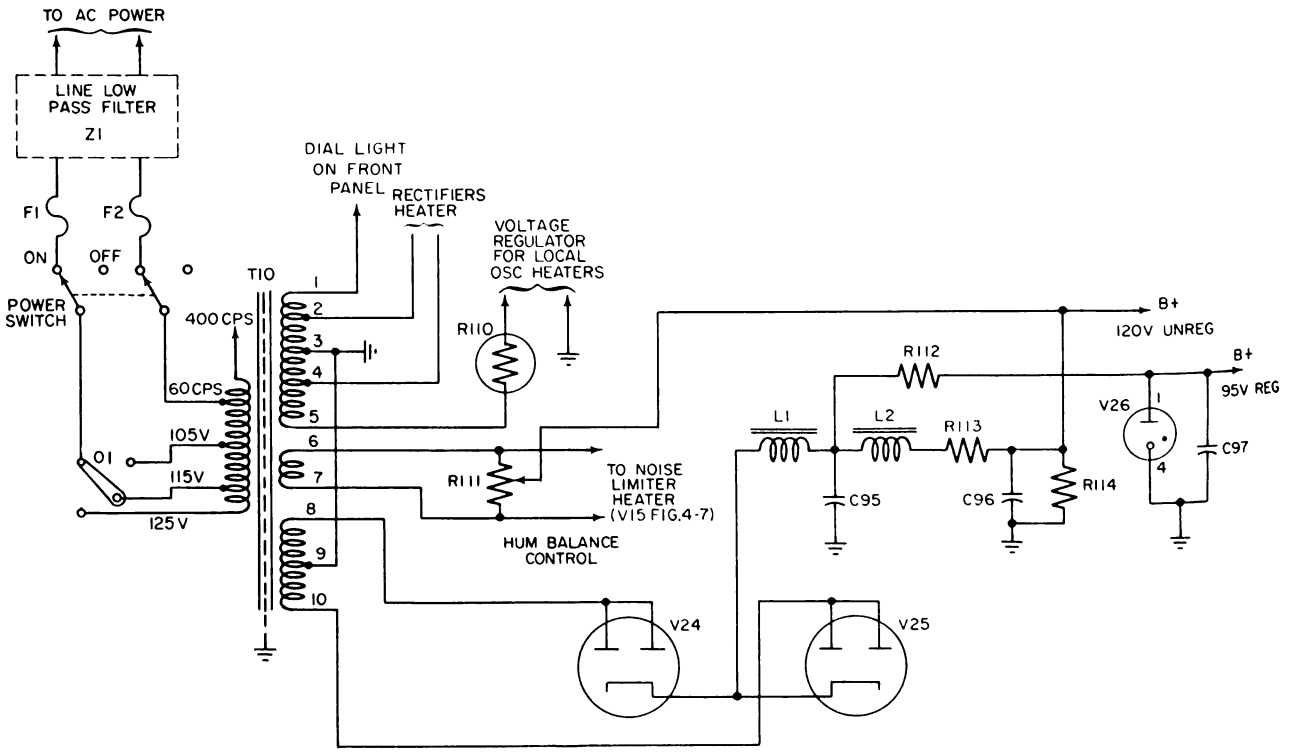

Figure 4-9. —Power supply for the AN/SRR-11, 12, and 13, receivers, simplified schematic.

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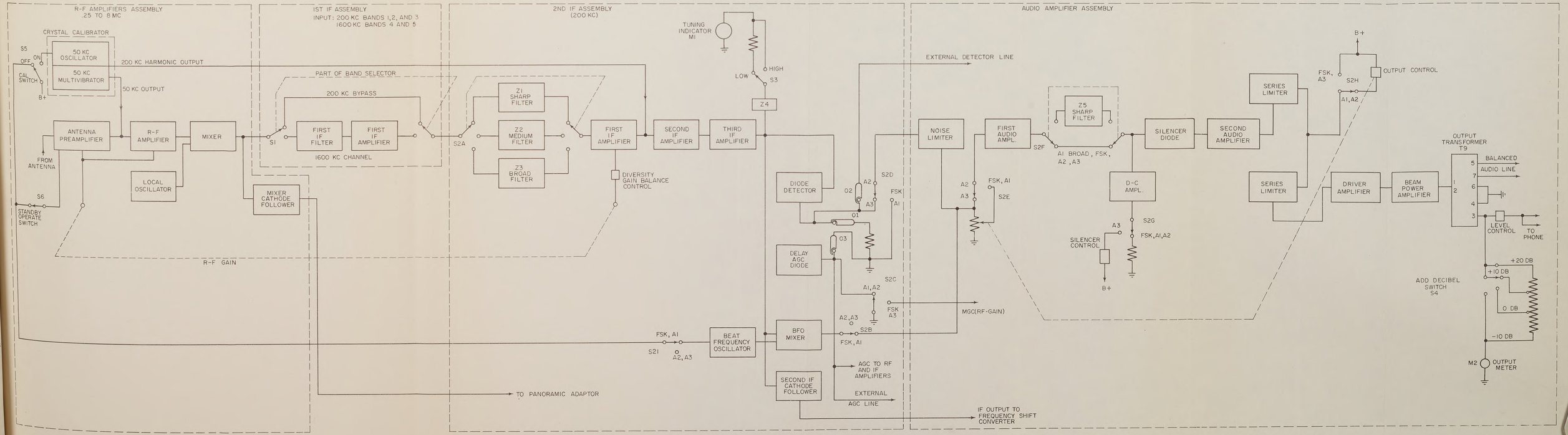

Figure 4-11. —Block diagram of Receiving set AN/SRR-12.

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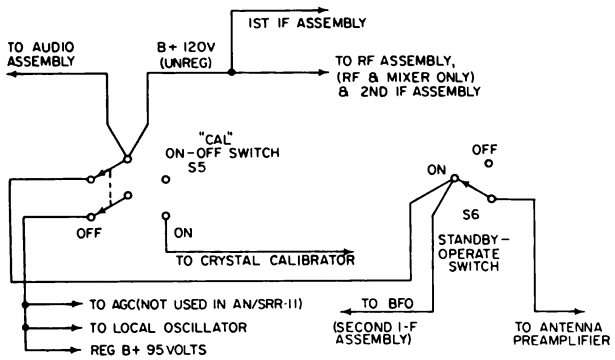

Figure 4-10. —Switches of the AN/SRR-11, 12, and 13 receivers, simplified schematic.

CIRCUITS OF THE AN/SRR-12 AND 13 RECEIVERS

The operation of the AN/SRR-12 and 13 receivers is similar to that of the AN/SRR-11 already discussed. For this reason, the discussion of the AN/SRR-12 and 13 receivers is limited to their major circuit differences as compared to the AN/SRR-11. A study of the block diagram of the AN/SRR-12 (fig. 4-11) and the AN/SRR-13 (fig. 4-12 pages 67 and 68) will show the similarity of the two receivers.

Crystal Calibrator

A crystal-controlled, cathode-coupled multivibrator (fig. 4-13) serves as the crystal calibrator in the AN/SRR-12 and 13 receivers. The circuits of the two receivers differ only in the relative component size.

The crystal, Y1, synchronizes (locks in) the multivibrator at 50 kc. C2 in parallel with the grid-to-plate capacity of the tube V1, increases the amount of feedback to the crystal. R2 is a parasitic suppressor. R6 is the common cathode resistor for both tubes. The ohmic value of R6 is low, and the resulting positive potential to ground is small.

Under static conditions (without synchronizing pulses from the crystal) V1 will conduct since its grid is tied to the cathode through R1. No current flows through R1 so that the grid-to-cathode potential of V1 is zero. Any tendency for conduction of V2 is counteracted by the voltage across R6 since the control grid of V2 is grounded through R5. For this reason, V1 plate potential with respect to ground is low, while the plate to ground potential of V2 is equal to the B supply value.

Negative pulses from the crystal are sufficient in amplitude to cut off V1, causing the plate potential with respect to ground to rise. C3 charges to the new value of plate voltage through R3 and R5, resulting in a positive going potential (with respect to ground) at the control grid of V2. The latter tube now conducts saturation current. The plate to ground potential of V2 (which drops because of the V2 conduction) will remain low for a time, depending on the charge of C3.

The cycle from the crystal then swings in the positive (or less negative) direction. When this occurs, the plate to ground potential of V1 drops, and C3 discharges through R5, making the V2 negative with respect to ground. V2 then cuts off, allowing its plate potential to return to the B supply value. The arrival of the next negative pulse from the crystal again cuts off V1, and the action recurs. In this manner, the multivibrator is synchronized to the 50-kc frequency of the crystal.

Crystal Calibrator Outputs

The multivibrator supplies two outputs, both of which are rich in harmonic content. The plate output is developed across C4 and R7. The output voltage from R7 is then coupled to the r-f amplifier of the associated receiver. The proper harmonic frequency of the first output is heterodyned in the mixer in accordance with the setting of the bandswitch and main tuning dial of the receiver. The desired 200-kc intermediate frequency is produced in the same manner as previously described in the discussion of the crystal calibrator of the AN/SRR-11. The second output, taken from the common cathode resistor R6, is coupled through C6 to the second i-f assembly where its fourth harmonic (200-kc) is used to produce the zero beat with the intermediate frequency during calibration of the main tuning dial.

The output of the r-f amplifier is coupled through the C section of S1C (a portion of the band selector switch) to the primary of the input transformer, T1 (fig. 4-14). The signal is coupled through the secondary of T1, via the A section of S1C, to the control grid of the mixer, V3. Fixed capacitors C7 and C10, trimmer C9 and the tuning capacitor, C5C, resonate with the secondary coil. C5C is shunted across the secondary of the selected transformer on

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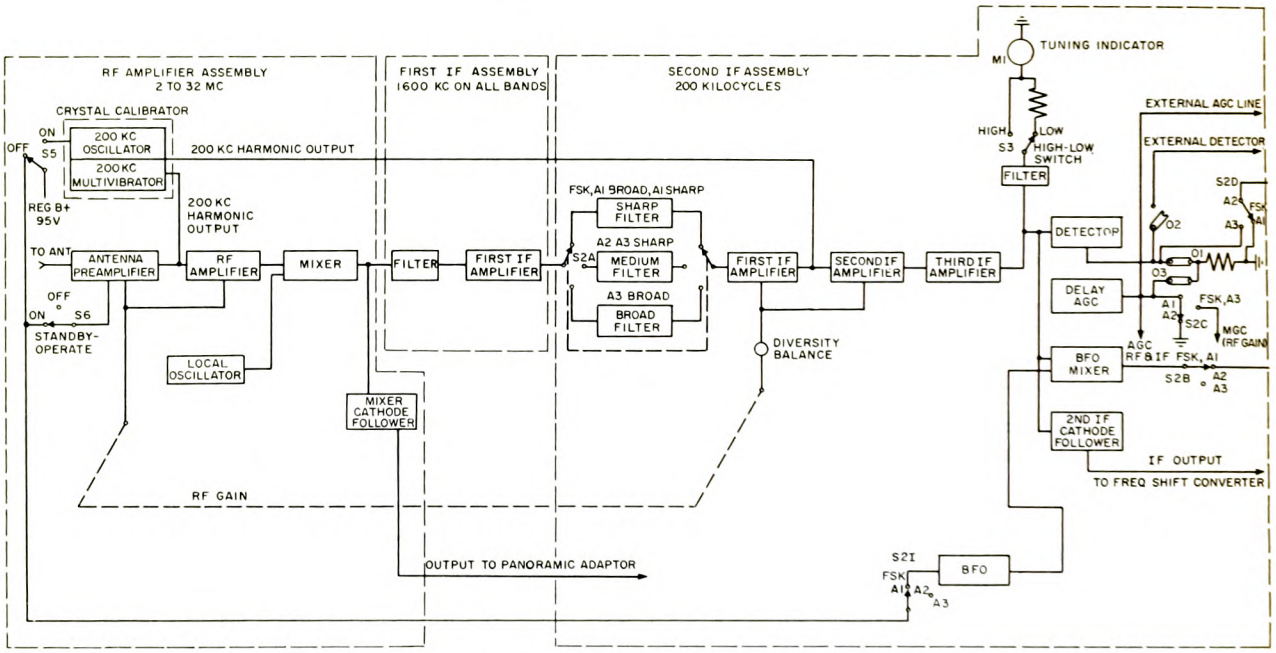

Figure 4-12. —Block diagram of Receiving Set AN/SRR-13.

all bands to provide tuning throughout the chosen band.

The electron-coupled Hartley oscillator, V4, has its cathode and control grid connected in the oscillatory circuit by band selector section S1D. The series cathode resistor, R45, provides a small cathode bias, which improves the stability of the oscillator. Plate and screen potentials are obtained from the regulated B supply to further insure frequency stability.

Signals from the local oscillator, V4, are applied to the suppressor grid of the mixer V3. The oscillator frequency tracks above the incoming frequency so that when the receiver is set on bands 1, 2, and 3, the output frequency of the mixer is 200 kc. The oscillator frequency on bands 4 and 5 is 1600 kc above the input frequency.

The output signals of either 200 or 1600 kc are applied along dual paths to the first i-f assembly and to the mixer cathode follower, V5.

In the AN/SRR-13 receiver, the oscillator tracks 1600 kc higher than the signal frequency on all five bands. Except for different frequencies, the mixer and oscillator circuits of the two receivers are the same.

Cathode Follower

A cathode follower, V5, (incorporated in the AN/SRR-12 and 13 receivers) provides an output for use with a panoramic adapter. Because of the input selectivity characteristics of the second i-f filters, the use of this output is recommended on bands 4 and 5 only. A detailed discussion of the operation of a panoramic adapter is given in chapter 9, *Electronic Technician* 2, Vol. 2, of this training course.

FIRST I-F ASSEMBLY

Filter

When the band selector of the AN/SRR-12 is set to the BAND 1 position (fig. 4-15), the signal input from the mixer (200 kc) passes through S1E and thence through S1F to the output transformer, T3. The filter, Z1, and converter stage, V6, are therefore bypassed. The same action takes place on bands 2 and 3 of the AN/SRR-12 when the mixer output frequency is again 200 kc.

When the first i-f input is 1600 kc (bands 4 and 5), the signal passes through the 4 (or 5) contacts of S1E to the filter, Z1. The filter passes a band of frequencies from 1585 to 1615 kc. This bandwidth is sufficient for passing

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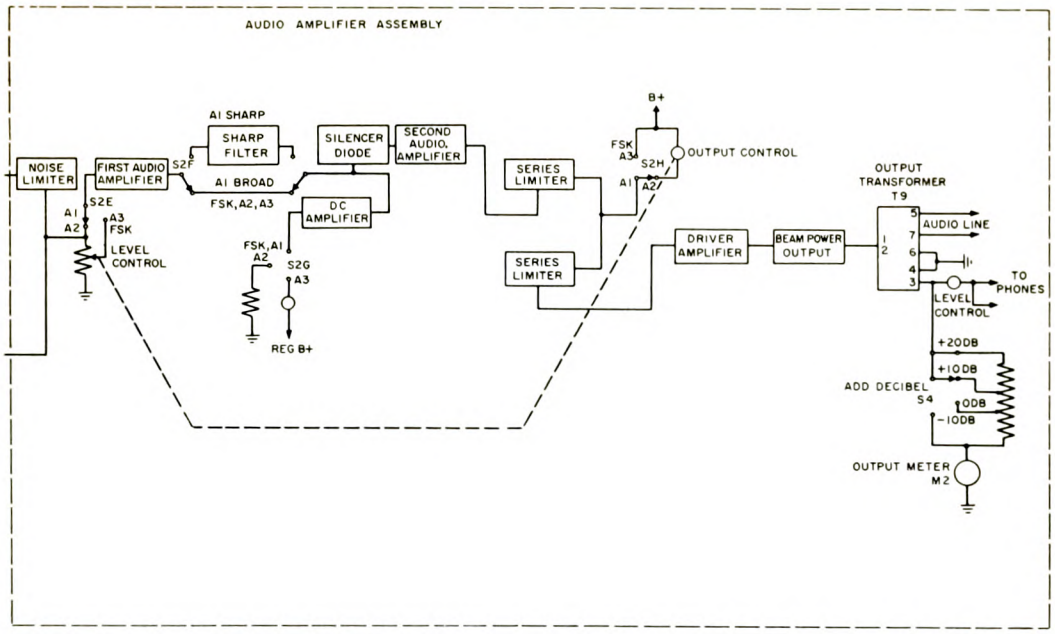

Figure 4-12. —Block diagram of Receiving Set AN/SRR-13 — Continued.

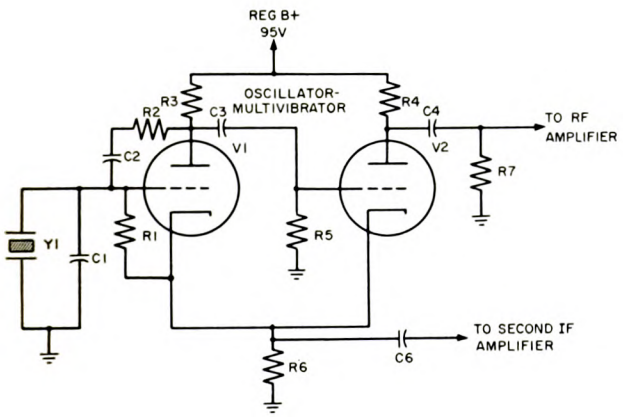

Figure 4-13. —Crystal Calibrator AN/SRR-12, 13 simplified schematic.

the signal intelligence while suppressing the undesired frequencies.

Converter

After passing through the filter, the 1600-kc signal is applied to the suppressor grid of the converter tube, V6. The oscillator section, which operates as a crystal-controlled Pierce circuit, comprises the control grid, cathode, and screen grid of tube V6. The oscillator frequency is 1400 kc. Feedback for the circuit is supplied by C30, C28, and through the interelectrode capacitance of the tube.

The 1400-kc signal produced by the oscillator is heterodyned in V6 with the 1600-kc signal from the filter to produce the desired 200-kc intermediate frequency. The output of V6 is fed through S1F (contact 4 or 5, depending on the band in use) to the primary of the output transformer, T3. The signal from the secondary of T3 is applied to the first i-f amplifier of the second i-f assembly.

The first i-f assembly of the AN/SRR-13 (not shown) does not contain band selector sections S1E and S1F since its input from the mixer is 1600 kc for all bands. Permanent connections are therefore made from the mixer output to the input of the band-pass filter. With this one exception, the circuit operation of the first i-f assembly of the AN/SRR-13 is the same as that discussed for the AN/SRR-12.

SECOND I-F ASSEMBLY

Filters and Input Circuits

The reception control, S2 (fig. 4-16) serves the same function in the AN/SRR-12 and 13, as previously described in the discussion of the AN/SRR-11. Since the AN/SRR-12 and 13 receivers are equipped for voice reception, the

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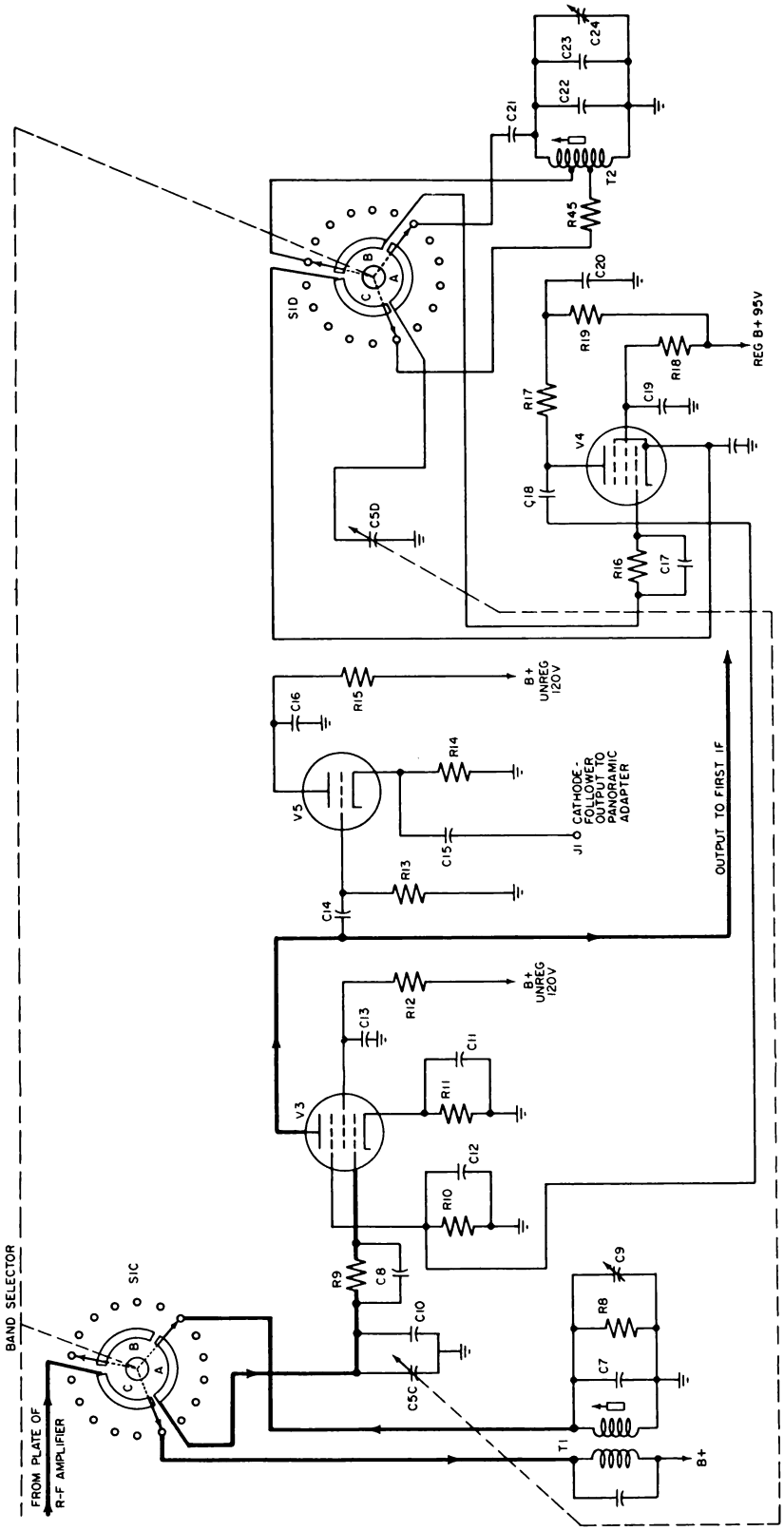

Figure 4-14. —Mixer, local oscillator, and mixer cathode follower of AN/SRR-12, simplified schematic.

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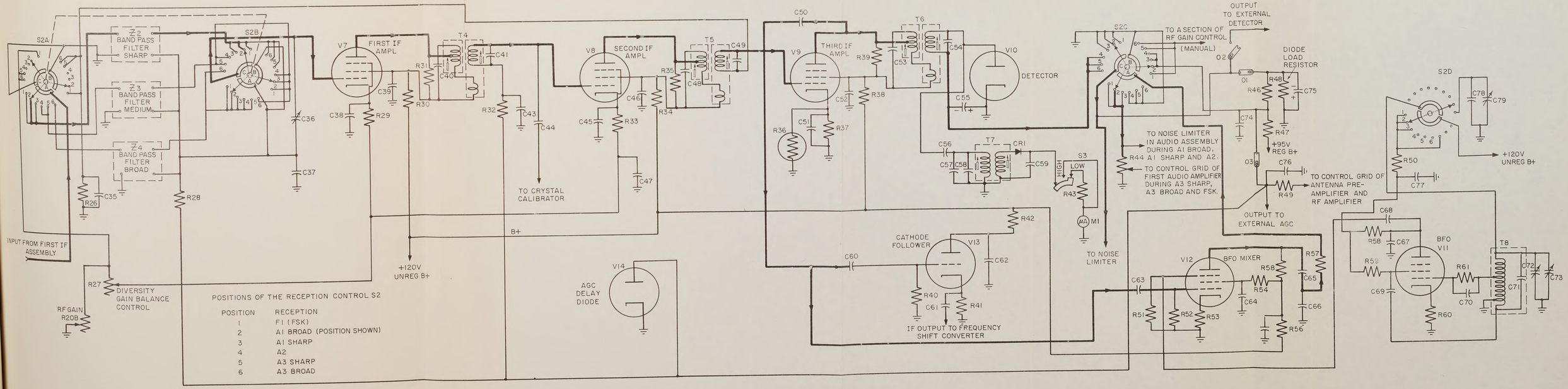

POSITIONS OF THE RECEPTION CONTROL S2

POSITION	RECEPTION
1	F1 (FSK)
2	A1 BROAD (POSITION SHOWN)
3	A1 SHARP
4	A2
5	A3 SHARP
6	A3 BROAD

Figure 4-16. —Second I-F Assembly AN/SRR-12 and 13, simplified schematic.

Figure 4-14. —Mixer, local oscillator, and mixer cathode schematic.

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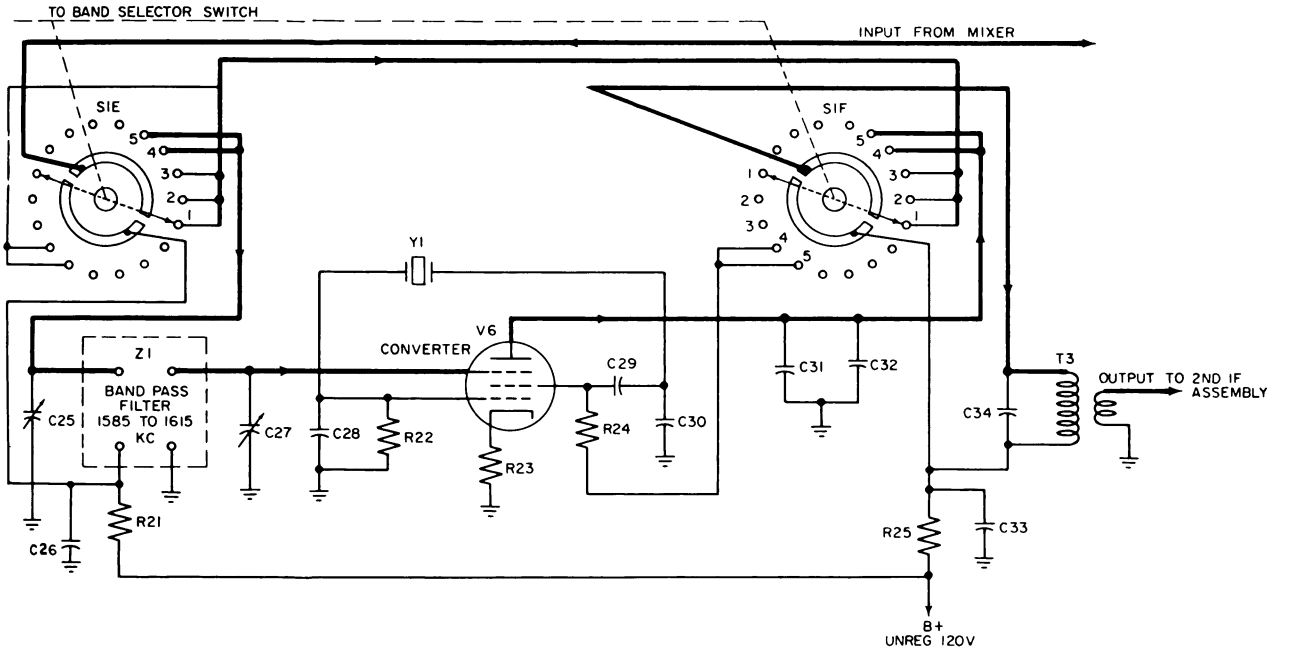

Figure 4-15. —First I-F Assembly AN/SRR-12, simplified schematic.

A3 BROAD and A3 SHARP positions have been added to the four positions of this switch. The switch sections of the reception control in figure 4-16 are shown in the A1 BROAD position.

The 200-kc input from the first i-f assembly passes through the A section of S2A to one of three filters, as chosen by the switch position. When the reception control is set to the FSK, A1 BROAD, or A1 SHARP position (corresponding for positions 1, 2, and 3) the "sharp" filter, Z2, is selected. During A2 and A3 "sharp" reception, the medium filter, Z3, is chosen (positions 4 and 5). A third filter, Z4, is chosen during A3 BROAD reception (position 6).

S2B accepts the signal from the output of the chosen filter and applies this signal to the control grid of the first i-f amplifier, V7. The operation of the i-f stages is the same as previously discussed.

Tuning Meter

The output of the third i-f amplifier, V9, is applied through C56 to a 200-kc tuned circuit comprising C57, C58, and the primary of T7. The secondary of T7 contains a crystal rectifier, CR1, which rectifies the secondary signal. The rectified current actuates the movement of the tuning meter, M1. C59 serves as a filter for the meter to eliminate spurious noise signals.

When the desired signal is detuned, the rectified signal voltage applied to the tuning meter is low, and the meter reads down scale. As the signal is properly tuned in, the rectified current increases, and the tuning meter reads up scale. The high-low switch, S3, functions in the same manner as previously discussed.

Detector and AGC Circuits

The 200-kc signal from the plate of V9 (fig. 4-16) is applied to the detector circuit, V10, through the broadly tuned transformer, T6. An audio voltage is developed across the diode load resistor, R48. C55 and C75 bypass r-f variations around the load resistor. The signal is applied to the audio noise limiter of the audio assembly when the reception control is set to the A2 or A3 positions (S2C positions 4, 5, and 6). With S2 set for A1 broad, A1 sharp and FSK reception (S2C positions 1, 2 and 3), the detector output is discontinued by the C section of S2C. During these modes of reception, signals from the BFO mixer, V12, are coupled to the control grid of the first audio

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amplifier by the A section of S2C. When the reception control is set to the FSK or A3 position (S2C positions 1, 5 and 6), the A section of the r-f (manual) gain control is shorted to ground by the B section of S2C.

The detector output voltage is also applied through R46, and link 03 (which is closed for single receiver operation) to the AGC delay diode, V14. C74 filters the modulation component of the detected signal so that only the d-c component is applied to the AGC diode. V14 is normally conducting as a result of a positive voltage (with respect to ground) which is applied from the +95-V regulated B supply. The AGC line therefore applies a slightly positive potential to ground to the control grids of the r-f and i-f amplifiers during no signal or weak signal condition.

The conducting path for V14 (fig. 4-17) is from ground, through the cathode-to-plate resistance of V14, and through R47 to the B supply. The network comprising R46, in series with the parallel combination of R48 and C75 is connected in parallel with the diode, and is effectively shorted by the low plate-to-cathode resistance of V14.

As the detector output signal increases, the drop across R47 increases, and the V14 plate is less positive (to ground). When the voltage across the diode load resistor, R48, exceeds a certain value (approximately 12 volts) V14 conduction ceases. Any further increase in signal voltage will reverse the polarity of the AGC output, making it more negative with stronger signals.

The negative voltage from the delay AGC diode is fed through R49 to the control grid of the antenna and r-f amplifiers, and from the plate of V14 to the control grid of the first and second i-f amplifiers of the second i-f assembly. The magnitude of the negative potential is determined by the value of the audit voltage in excess of the cutoff potential of V14.

The action of the B section of S2C is such that delayed AGC is effective during F1 and A3 reception of the AN/SRR-12 and 13. In these modes of reception, the A section of the r-f gain control is shorted to ground through contacts 1, 5, or 6 of the S2C switch section B. When S2C is set (by the reception control) to the A1 BROAD, A1 SHARP, or A2 positions, the delayed AGC line is grounded through contacts 2, 3, or 4 of section B. The receiver gain is then controlled by the A and B sections of the r-f gain control.

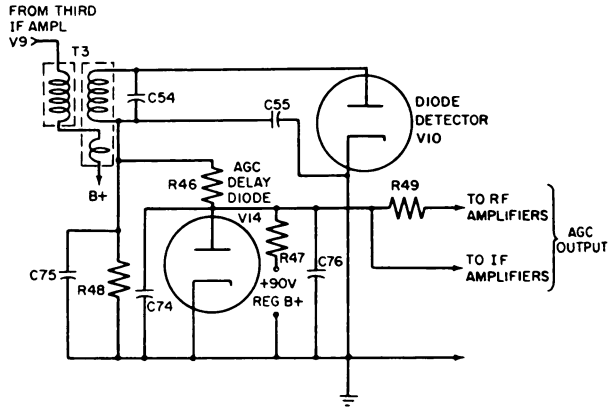

Figure 4-17. —Detector and AGC circuit, simplified schematic.

Dual Diversity Reception

Links are provided in the diode detector and AGC circuits of the AN/SRR-12 and 13 receivers for receiving the detector rectified signal voltage from another receiver, and for applying the automatic gain control voltage to a second AN/SRR-12 or 13 receiver during diversity reception. The receiver that supplies the AGC output is referred to in this discussion as the master receiver, while the unit receiving this voltage is designated the slave receiver.

The links, 01, 02, and 03, of figure 4-16 are shown in their respective positions for single receiver operation. If this receiver were used as the master receiver during dual diversity reception, all of the links would be in the closed position. In the slave receiver, only link 02 is closed.

The master receiver develops the AGC potential, which is then applied to the slave receiver as shown in figure 4-18. Since link 01 in the slave receiver is open, its audio section is not used. The output of both receivers is developed across R48.

If the signal from the master receiver is the stronger of the two, the AGC voltage to the slave receiver will be large enough to cut down its gain. This reduces the total AGC voltage and allows the master receiver to supply its normal output. If the signal from the slave receiver were of greater magnitude, the AGC voltage would act to decrease the gain of the master. Likewise then, the slave receiver is supplied with a smaller AGC voltage, such that its gain increases to normal value. If the signals

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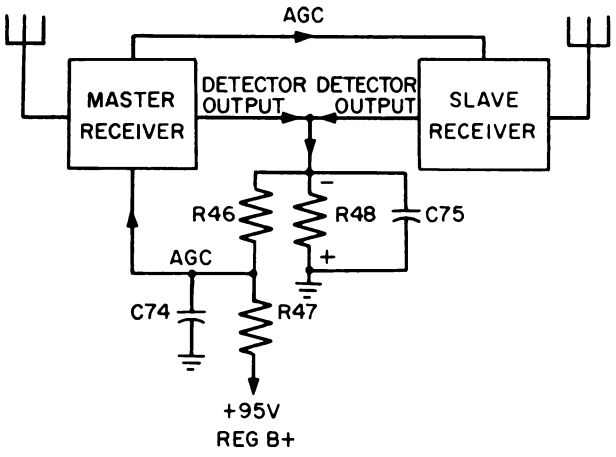

Figure 4-18. —Receiver connections for diversity reception, simplified diagram.

from both receivers were of equal magnitude, the AGC voltage would be twice its normal value, and the gain of each receiver is proportionately reduced.

The B section of the r-f (manual) gain control, R20B (fig. 4-16) is connected in series with the cathode resistors, R29 and R33, of the first and second i-f amplifiers to control the gain of the stages. The diversity gain balance control, R27, is also connected in series with the cathode resistors of these amplifiers. R27 is used for matching the gain of the master and slave receivers during diversity reception.

AUDIO ASSEMBLY

Silencer Circuit

A silencer circuit comprising tubes V17 and V16 (fig. 4-19) is active in the audio assembly of the AN/SRR-12 and 13 receivers when the reception control (section S2E) is set to the A3 SHARP or A3 BROAD position. The presence of the silencer circuit eliminates noise at the output of the receivers in the absence of signals.

The silencer diode, V16 is connected in series between the first and second audio amplifiers. The plate to ground potential of V16 is determined by the amount of plate current flowing through the d-c amplifier, V17. A voltage divider comprising R73 and R74 is connected between +125 volts and ground. The potential on this divider from the R73, R74 junction to ground supplies a fixed voltage for the V16 cathode.

The d-c amplifier, V17, receives a control grid potential, which is positive to ground by an amount determined by the setting of the silencer control, R63. R63 is connected between the +95 volts B+ and ground and thus supplies an adjustable positive voltage. The potential to the V17 grid is applied through R62 and section S2E of the reception control (in A3 sharp and A3 broad reception). R62 increases the resistance in the grid-to-cathode circuit so that the grid-to-cathode voltage does not become more than a few volts positive. In addition, R62 serves as a minimum resistance in the control grid circuit so that the V17 grid cannot be shorted to ground when the arm of R63 is moved to the ground position.

When the arm of R63 is moved above ground the grid voltage of the d-c amplifier, V17, is set to a certain positive value (within the limit of R63). This action increases the V17 plate current and lowers the V17 plate voltage so that the V16 plate becomes negative with respect to its cathode. Therefore V16 cannot conduct.

The audio signal input from the detector should be recalled as a negative voltage with respect to ground. When audio signals are received, the control grid-to-ground potential of V17 is reduced (less positive), and its plate voltage rises. V16 can therefore conduct. Audio input signals are filtered by C80 and R64 to provide a relatively constant input potential to V17, and to prevent noise signals from triggering the circuit.

Silencer Circuit Output

Since V16 is connected in series with the output signals, its action is similar to that of a switch. When V16 conducts (signals being received) the audio output from V16 is coupled by C84 and R75 (in parallel with the cathode to ground resistance) to the control grid of the second audio amplifier. When V16 does not conduct (corresponding to no signal input), the signal path to the second audio amplifier is discontinued.

EQUIPMENT CHANGES

The following changes have been made on the AN/SRR-11 receivers bearing serial numbers from 1276 and above.

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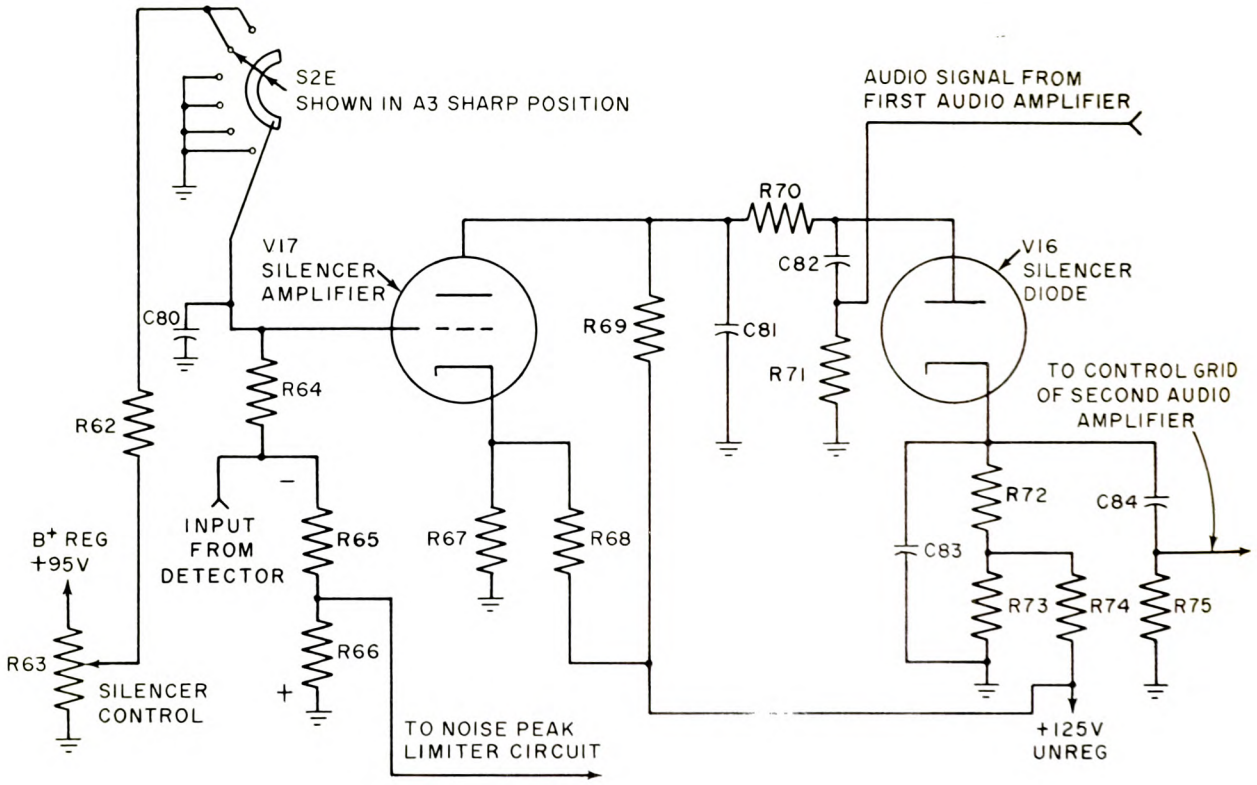

Figure 4-19. —Silencer circuit, simplified schematic.

Five of the type 5647 tubes (one in the second i-f assembly and four in the audio assembly) have been replaced with type 1N458 crystal diodes. The changes are as shown in table 4-4.

Plug-in units having electron tubes are interchangeable with plug-in units having crystal diodes. The crystal diodes and tubes are also interchangeable.

Table 4-4. —Equipment Changes.

Tube symbol	Figure No.	Tube function	Major assembly
V9	4-6	Detector	Second i-f
V15	4-7	Noise limiter	Audio
V18	4-7	Limiter	Audio
V19	4-7	Limiter	Audio
V16	4-19	Silencer	Audio

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QUIZ

1. Transmission of frequencies in the 300- to 3000-kc range usually employs the use of what type waves for propagation?
2. Commercial broadcast as well as some Navy stations employ the use of frequencies in what band?
3. Frequencies in the 30- to 300-mc range yield a particular advantage over lower frequencies due to what feature of the antenna?
4. Frequencies greater than 30 mc usually suffer great losses caused by what earth constant?
5. (a) The AN/SRR-11, 12 and 13 cover what frequency range? (b) Each of the receivers has its total frequency range divided into how many bands?
6. Signals coupled from the antenna (fig. 4-3) are amplified in how many stages of r-f amplification?
7. (a) On bands 1 and 4 (fig. 4-3), what is the mixer output frequency? (b) On bands 2, 3, and 5, what is the output frequency of the mixer?
8. When the output frequency of the mixer is 60 kc, the switch, S1 (fig. 4-3) is set so that the signal passes through the first i-f assembly through what path?
9. The first i-f amplifier of the first i-f assembly (fig. 4-3) may best be described as what type circuit?
10. The crystal calibrator circuit provides accurate frequency check points that may be used for what purpose?
11. What switch (fig. 4-4) permits the choice of bands in accordance with the frequency it is desired to receive?
12. Continuous tuning within the chosen band is accomplished through the use of the ganged sections of what capacitor?
13. Both the input and output transformers in what section of the receiver are changed for each respective band?
14. The antenna used with the AN/SRR-11 aboard ship is usually of what type?
15. The antenna section of the ganged tuning capacitor, C5A (fig. 4-4) is shunted across the secondary of the antenna transformer through what section of the band selector, S1?
16. L1, in series with the r-f gain control serves as a filter for what purpose?
17. (a) The oscillator V4 (fig. 4-4) used to produce the heterodyning frequency comprises what type circuit?
    (b) Feedback in an amount sufficient to sustain oscillations, is provided by what capacitor?
18. (a) The oscillator (fig. 4-4) tracks on the high side of the signal input so that its output is what frequency above the input frequency on bands 1 and 4? (b) How much above the incoming frequency is the oscillator output on bands 2, 3, and 5?
19. The oscillator (fig. 4-5) used to produce the 140-kc heterodyning frequency is connected as what type circuit?
20. The characteristics of the 60-kc filter, Z1 (fig. 4-5) are such that they allow the passage of a band of frequencies between what limits?
21. When the reception control (fig. 4-6) is set to the FSK, A1 BROAD, or A1 SHARP positions, the output from the first i-f assembly passes through the A section of switch S2 via which contacts for each of the modes of reception?
22. (a) The A section of the r-f gain control, R20A (fig. 4-4) controls the gain of what stages? (b) The B section of the control, R20B (fig. 4-6) is connected in series with the cathodes of what stages?
23. R51, in parallel with the biasing network of the third i-f amplifier (fig. 4-6) is what type resistor?
24. The best frequency oscillator comprising V10 (fig. 4-6) provides the heterodyning signal to the BFO mixer, V11, when the reception control is placed in what positions?
25. The frequency vernier on the receiver front panel controls what component of figure 4-6?
26. Input to the audio assembly (fig. 4-7) is determined by the setting of what control?
27. When a noise pulse (or modulated signal that exceeds normal modulation percentage) is detected, what effect will this have on the noise limiter diode (fig. 4-7)?
28. The signal to the output meter, M2 (fig. 4-7) may be decreased in four steps by what switch in the audio assembly?
29. The 10-kc crystal calibrator (fig. 4-8) comprises what two circuits?
30. (a) What components (fig. 4-8) serve as the main frequency determining components of the oscillator? (b) The feedback for the oscillator is coupled through what component?
31. The 10-kc multivibrator, V23 (fig. 4-8) is synchronized from what source?
32. (a) The 10-kc multivibrator output (fig. 4-8) is coupled from the cathode of V23 to what unit? (b) The fourth harmonic of the 50-kc output (fig. 4-8) is taken from the oscillator plate and applied to what unit?

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33. A regulated +95 volts is obtained from what source in the power supply circuit (fig. 4-9)?
34. The cal switch, S5 (fig. 4-10) is connected so that in the OFF position the B supply voltage is not applied to what circuits?
35. What stage serves as the crystal calibrator in the AN/SRR-12 and 13 receivers (fig. 4-13)?
36. The electron-coupled Hartley oscillator (fig. 4-14) has its cathode and control grid connected in the oscillatory circuit through the action of what control?
37. In the AN/SRR-13 receiver the oscillator tracks 1600 kc higher than the signal input frequency on what bands?
38. When the reception control is set to the A1 BROAD, A1 SHARP, or FSK position (corresponding to positions 1, 2, and 3 of S2A in figure 4-16,) what input filter is selected?
39. The detector output voltage (fig. 4-16) is applied through R46, and link 03 (which is closed during single receiver operation) to what tubes in the second i-f assembly?
40. (a) The negative voltage from the delay AGC diode (fig. 4-16) is fed through R49 to the control grid of what stages? (b) The signal is coupled from the plate of V13 to the control grids of what stages in the second i-f assembly?
41. The action of the B section of S2C (fig. 4-16) is such that delayed AGC is effective during what modes of reception in the AN/SRR-12 and 13?
42. During dual diversity reception, the master receiver (fig. 4-18) develops the AGC potential, which is then applied to what unit?
43. A silencer circuit (fig. 4-19) comprising tubes V14 and V16 is active in the audio assembly of the AN/SRR-12 and 13 receivers when the reception control (section S2E) is set to what position?

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CHAPTER 5

OPERATING PRINCIPLES OF A REPRESENTATIVE RADIO TRANSMITTER, PART I

INTRODUCTION

The radio frequency oscillator of the transmitting sets AN/SRT-14, 15, and 16 constitutes the major part of the discussion of this chapter. The radio frequency oscillator is the master frequency source of the transmitter for all modes of operation.

In order to provide its wide variety of frequencies, the radio frequency oscillator is divided into subunits, each of which perform a definite function in the generation of the desired output frequency. Some of the subunits are frequency converters or mixers, while others are frequency multipliers. Cathode followers and oscillators make up other subunits.

In many transmitters when it is desired to change the output frequency (for example, from 5 to 10 megacycles) the oscillator frequency must be changed accordingly. Although this principle is widely used and accepted, it is somewhat inefficient. This is true since each change in the oscillator operating frequency affects the circuit stability of that oscillator.

The oscillator used in the radio frequency oscillator of the AN/SRT-14, 15, and 16 (during c-w and phone operation) has a fixed frequency of 100 kc and provides a transmitter output frequency ranging from 0.3 to 26 mc. These output frequencies are made possible through heterodyning processes taking place in several subunits throughout the radio frequency oscillator. This process is relatively new in transmitter circuitry. The mixing circuits are therefore made variable in preference to the circuitry of the oscillator. Since the oscillator output is fixed at one frequency (100 kc), it provides a more stable output to the other subunits of the radio frequency oscillator.

Any one of the transmitting sets, AN/SRT-14, 15, or 16, is capable of performing the operations that heretofore has required the use of several separate transmitters. This fact alone makes it easy to see why the AN/SRT-14, 15, 16 are rapidly replacing such transmitters as the TBL, TBK, TBM, and TAJ. Further, all of the major components of each transmitter are mounted on separate chassis, which are contained in a single cabinet approximately 55 in. high and 16 in. wide. Therefore, the space used by the newer transmitters is considerably less than that of the older models, such as the TBM, which employs the use of a separate cabinet containing the modulator.

The circuits contained in the radio frequency oscillator are generally simple, and, in many cases, representative of circuits you have studied in the past and will encounter in the future. The circuits are selected so as to overcome many of the disadvantages of earlier transmitters. Jacks, meters, and indicator lamps are incorporated throughout the transmitter, which make possible visual checks for proper circuit operation, or, which may be used as aids in servicing.

The order of presentation is (1) a description of the capabilities of the three transmitters, (2) a block diagram analysis of the radio frequency oscillator common to all three of these units, and (3) a circuit analysis of each stage of the radio frequency oscillator.

RADIO TRANSMITTING SETS AN/SRT 14, 15 AND 16

Radio Transmitting Sets AN/SRT-14, 15, and 16 provide radio-frequency energy at any frequency from 0.3 to 26 mc for the transmission of voice, facsimile, and also c-w or frequency shift telegraph communication. A manual frequency selection process permits all three sets to cover the entire frequency range in steps of 10 cycles at the 100-watt output level. The AN/SRT-15 and AN/SRT-16 can also operate at the 500-watt carrier level in the frequency range from 2 to 26 mc.

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A pictorial representation of the AN/SRT-14 with its major components is shown in figure 5-1. The set consists of a 100-watt transmitter bay and the antenna tuning equipment. The Radio Transmitting Set AN/SRT-15, shown in figure 5-2, consists of a transmitter bay and booster and the antenna tuning equipment. The booster provides the necessary facilities for producing the 500-watt operating output level. Therefore, the AN/SRT-15 may transmit an r-f carrier of either 100 or 500 watts. Although the carrier levels are referred to as 100 and 500 watts, the output varies over the frequency range according to the frequency and to the form of communication.

The Radio Transmitting Set AN/SRT-16 (fig. 5-3) consists of two transmitter bays, each of which is equipped with its antenna tuning units. The dual transmitter is therefore capable of transmitting two carriers simultaneously. One of the outputs is limited to the 100-watt level, while the other may be transmitted at either 100 or 500 watts.

Either of the transmitter sets may be arranged for remote operation with start-stop control, keying, and radiophone. Standard Navy

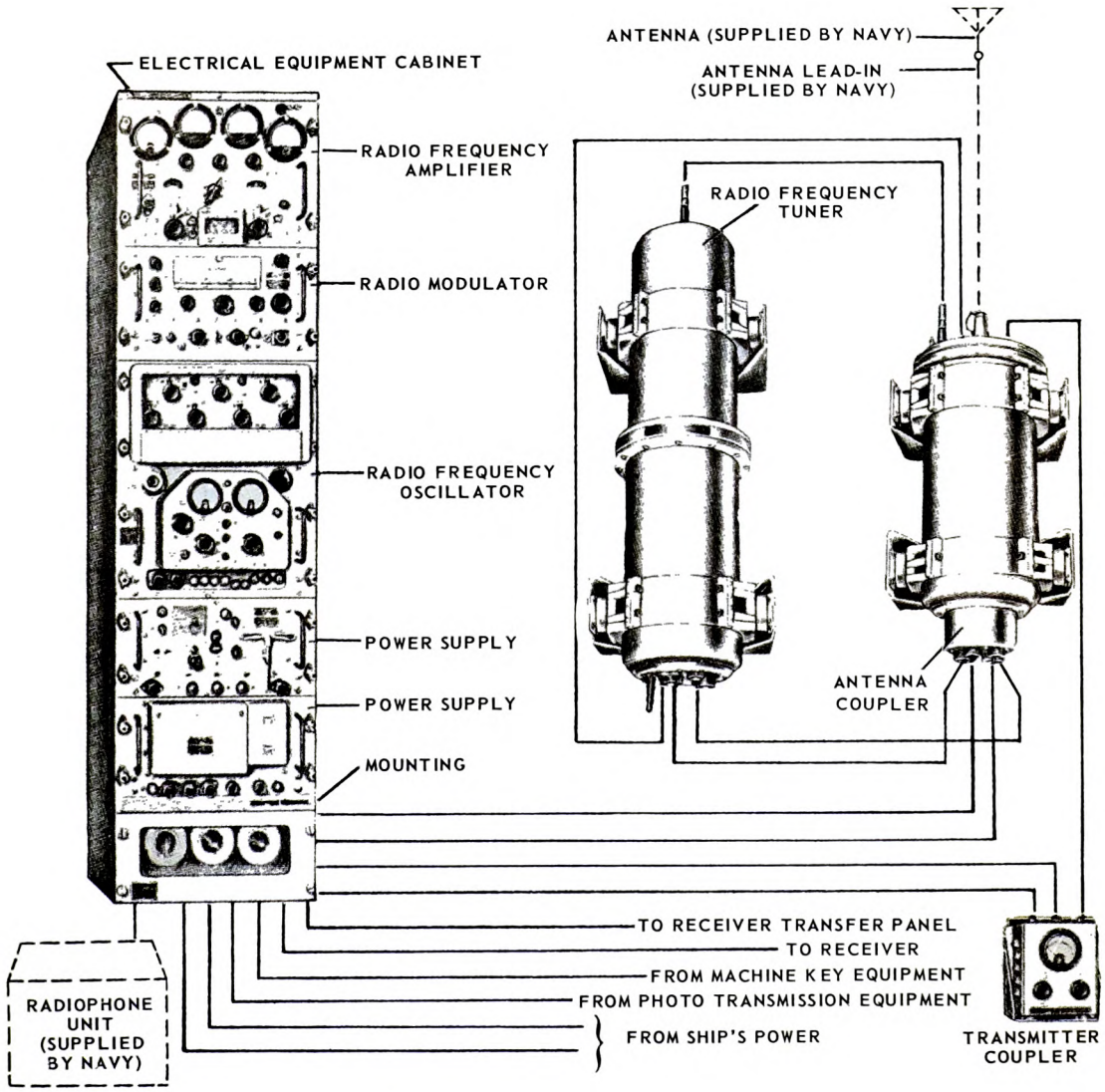

Figure 5-1. —Radio Transmitting Set AN/SRT-14, relationship of units.

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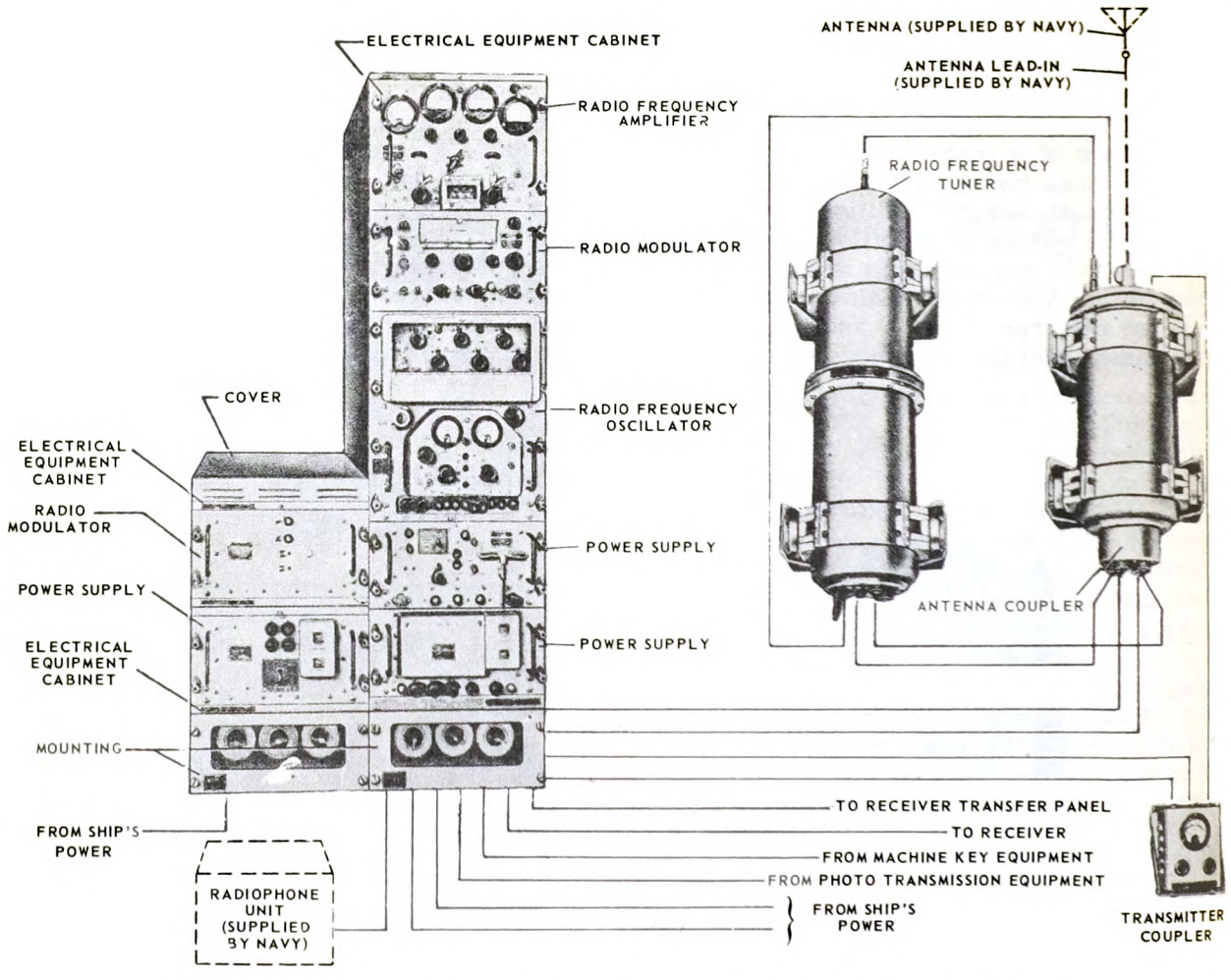

Figure 5-2. —Radio Transmitting Set AN/SRT-15, relationship of units.

six-wire radio transmitter control circuits are used for all remote control operations.

Because of the similarity of the three transmitters, a single functional block diagram is shown in figure 5-4. The diagram is used to show the relationship existing between the major units and the progression of the signal through the set. A full description of each major component of the transmitter is given later in the chapter.

The radio frequency oscillator is the master frequency source for the transmitting set. It generates its output signals through the use of 14 subunits, employing oscillators, frequency multipliers, and frequency converters. The r-f output of the radio frequency oscillator (approximately two volts) can be varied in steps of 10 cps from 0.3 to 26 mc. This output is fed to the radio frequency amplifier. The latter circuit contains three stages of amplification; a buffer, an intermediate power amplifier, and a power amplifier. These stages are manually tuned to the desired frequency. During 100-watt operation, the radio frequency amplifier amplifies all frequencies in the 0.3 to 26-mc range. In 500-watt operation, only signals within the range from 2 to 26 mc are amplified in the radio frequency amplifier. Selection of a frequency between 0.3 and 2 mc during 500-watt operation automatically switches the output to the 100-watt level.

The low-level radio modulator accepts audio and keying signals from the external circuits, and amplifies and shapes them as required to modulate the carrier. A 50-watt audio signal is directly coupled to the radio frequency amplifier

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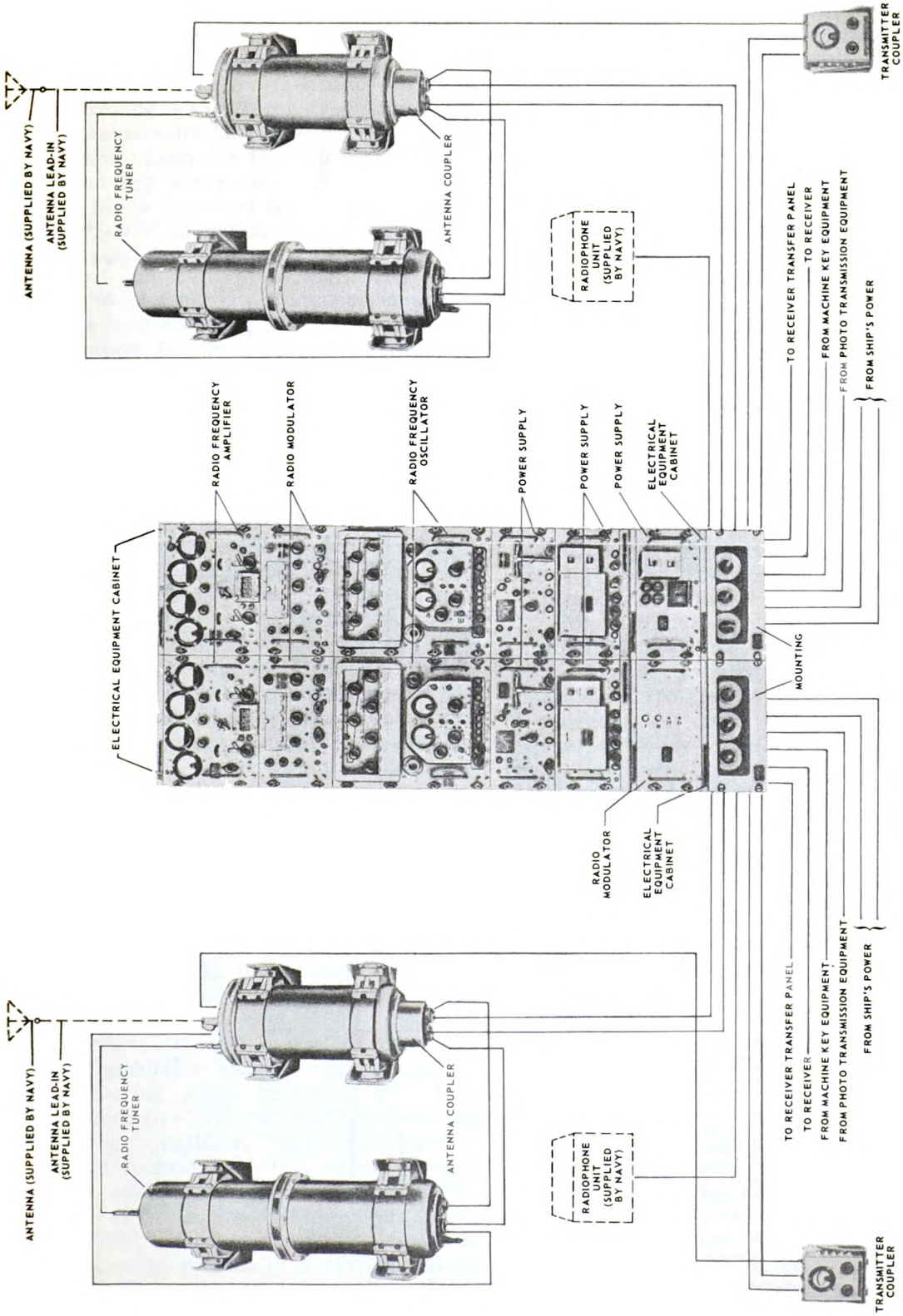

Figure 5-3. —Radio Transmitting Set AN/SRT-16, relationship of units.

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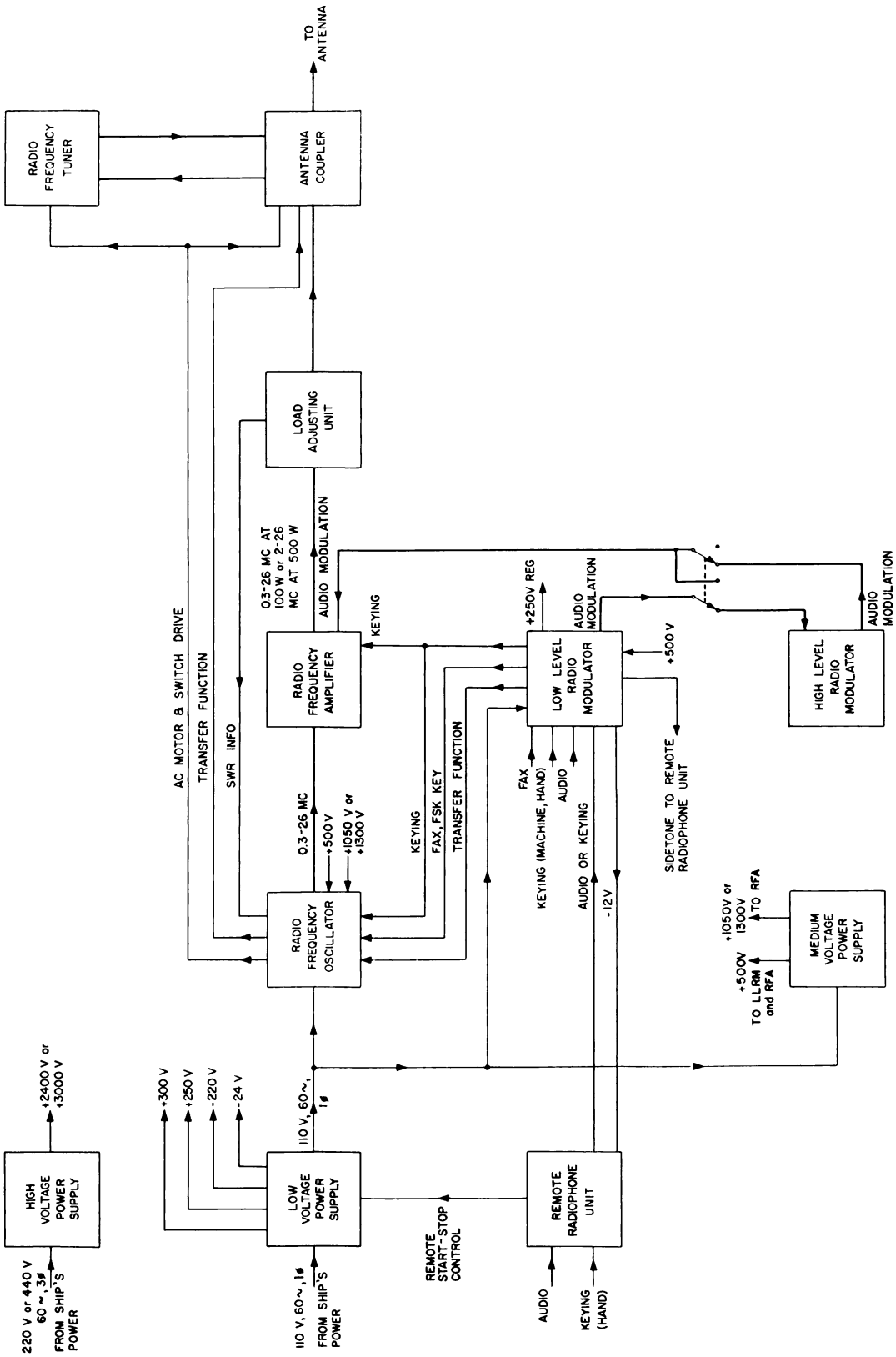

Figure 5-4. —Functional block diagram of Radio Transmitting Sets AN/SRT 14, 15, and 16.

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during 100-watt operation. During 500-watt operation, the low-level radio modulator supplies a 6-watt audio signal to the high-level radio modulator, which amplifies this input to the 250-watt level and feeds it to the radio frequency amplifier. Keying signals for frequency-shift and c-w transmission may be applied to the low-level radio modulator from either machine-key or hand-key equipment. The high-level modulator, which provides the required audio level for 500-watt operation is used only in the AN/SRT-15 and 16.

The output of the radio frequency amplifier is fed to the load adjusting unit. This unit provides a method for improving the impedance match between the output of the radio frequency amplifier and the antenna. From the load adjusting unit, the signal is applied to the antenna coupler whose prime function is to extend the range of tuning. This is done by switching inductance or capacitance into the line to lengthen or shorten the apparent electrical length of the antenna. The antenna coupler also contains a switch, which makes it possible to completely bypass the tuning components, thereby connecting the antenna directly to the load adjusting unit. The radio frequency tuner serves as a variable length transmission line. Its function is to sufficiently match the impedance of the antenna to the characteristic output impedance of the radio frequency amplifier so that the standing wave ratio will be no greater than 4:1.

The low-voltage power supply receives 110 volts a-c from the ship's supply and distributes this power to other units. It contains rectifier circuits that provide B+ voltage of +300 volts, +250 volts, -24 volts for control circuits and motor lines, and -220 volts for bias.

The medium-voltage power supply receives 110 volts from the low-voltage power source, and supplies +500 volts and +1300 volts, respectively. The +500 volts is applied to the low-level radio modulator and the radio frequency amplifier. During phone operation, the +1300 volts is reduced to +1050 volts and applied to the radio frequency amplifier. The medium-voltage power supply contains a relay, which ensures that the 500-volt source is activated before the 300-volt source in the low-voltage power supply comes on. This action is necessary since the 300 volt source is primarily a screen voltage supply and will cause damage to the tubes it serves if plate voltage has not been applied.

The high-voltage power supply used only in the AN/SRT-15 and 16 during 500-watt operation is provided with a separate input of 220 volt, single phase or 440 volt, three phase, 60 cycle, a-c from the ship's supply. This circuit has a d-c output of +3000 volts that replaces the 1300-volt output of the medium voltage power supply during 500-watt operation. During phone operation, the 3000 volts is reduced to +2400 volts.

As previously stated, the types of signals that may be applied to the input circuits are voice, c-w, frequency-shift telegraph (hand or machine keyed), or facsimile. The radiophone unit (not a part of the radio transmitting sets) is usually located at a remote position. This unit consists of a power start-stop circuit and an audio and keying input circuit. The power start-stop circuit is connected to the control circuit of the low-voltage power supply, and may be used to turn the transmitter on or off. Audio signals are received locally from either a carbon or dynamic microphone. When a standard Navy remote radiophone unit is used, the input circuit is restricted to the use of a carbon microphone.

For c-w operation, the low-level radio modulator feeds keying signals directly to the radio frequency amplifier and radio frequency oscillator at either 100- or 500-watt levels. In frequency-shift key operation, the low-level radio modulator reshapes the input keying signals, and feeds them to the radio frequency oscillator. Facsimile signals are switched through the low-level radio modulator and applied directly to the radio frequency oscillator.

Over the entire frequency range, the unmodulated carrier level for radiophone communication may vary from 55 to 100 watts at the low level, and from 265 to 400 watts at the higher level. The same frequency range carrier level, however, for telegraph and facsimile transmission may vary from 85 to 150 watts for the lower output level, and from 400 to 600 watts for the higher output level.

In the discussion of the AN/SRT-14, 15, and 16 in this chapter, it is sometimes best to use abbreviations for major components of the sets. These major components and abbreviations are:

Radio Frequency Oscillator (RFO)
Low-Voltage Power Supply (LVPS)
Medium-Voltage Power Supply (MVPS)
High-Voltage Power Supply (HVPS)
Low-Level Radio Modulator (LLRM)
High-Level Radio Modulator (HLRM)

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Radio Frequency Amplifier (RFA)
Antenna Tuning Unit (ATU)
Load Adjusting Unit (LAU)

RADIO FREQUENCY OSCILLATOR

As previously stated, the radio frequency oscillator is the master frequency source for the transmitting set. It contains three individual oscillators and several supporting units including frequency converters, step generators, and frequency multipliers, which are necessary to produce the required frequencies for the transmitter. The radio frequency oscillator unit, therefore consists of 14 subunits, which in this discussion, will be referred to by their common name and unit number, as shown in the radio frequency oscillator block diagram of figure 5-5, A. Figure 5-5, B, shows an external view of the radio frequency oscillator control panel.

DISCUSSION OF BLOCK DIAGRAM

Unit 1

The crystal oscillator (unit 1) produces a 100 kc output, which is fed to a frequency multiplier (unit 2), a frequency converter (unit 5), the 10-kc step generator (unit 6), and the 100-kc step generator (unit 8). This output synchronizes the step generators, and through frequency multipliers, provides higher frequencies as required by other units of the radio frequency oscillator. Another portion of the 100-kc output is fed to a zero adjust indicating circuit in unit 14 (front panel of fig. 5-1) to enable the operator to check the output of the interpolation and frequency-shift oscillators. Finally, a portion of the 100-kc output of the crystal oscillator is fed to an oscilloscope test receptacle in unit 14 for servicing or checking as required.

Unit 2

In the frequency multiplier (unit 2), the 100-kc output of the crystal oscillator is increased to 1 mc and fed to another frequency multiplier (unit 4), a frequency converter (unit 5), a 10-kc step generator (unit 6), a frequency multiplier (unit 7), and a 1-mc step generator (unit 10). Unit 4 multiplies the 1-mc input frequency from unit 2 and produces an 8-mc output, which is fed to the frequency converter (unit 5).

Unit 5

In addition to the 1-mc input from unit 2 and the 8-mc input from unit 4, the frequency converter (unit 5) receives outputs from the interpolation oscillator (unit 3) and either the crystal oscillator or the frequency-shift oscillator. The interpolation oscillator produces an output of 90 to 100 kc, adjustable in steps of 10, 100, and 1000 cps. The major portion of this output is fed to the frequency converter (unit 5). Another portion is fed to an oscilloscope test receptacle located on the front panel of the RFO. The frequency-shift oscillator (unit 12) generates an output of 100 kc, frequency modulated in accordance with signals applied from the low-level radio modulator. For frequency-shift keying operations, the output may be shifted a maximum of ±1000 cps or $\Delta f/2$ from the knob set frequency. For facsimile operation, the output frequency shift may be adjusted from +500 to +2000 cps (or Δf) for a positive 20-volt signal from the facsimile equipment. During either frequency-shift keying or facsimile operation, the output of unit 12 is fed to the frequency converter (unit 5) to replace the 100-kc signal normally supplied by the crystal oscillator (unit 1). Another portion of the frequency-shift oscillator output is fed to an oscilloscope test receptacle. The frequency converter (unit 5) consists of a series of conventional mixers, tuned filters, and an amplifier to heterodyne its four input frequencies. The output of unit 5 is adjustable in steps of 10 cps from 9.19 to 9.2 mc for c-w and phone, or 9.19 to 9.2 mc ±1000 cps (maximum) for frequency-shift telegraphy, and 9.19 to 9.2 mc +2000 cps (maximum) for facsimile operation.

Unit 6

The output of unit 5 is fed to the 10-kc step generator (unit 6) whose major component is a phase-locked oscillator. The 10-kc step generator receives, in addition, a 100-kc signal from the crystal oscillator and a 1-mc signal from unit 2. The 100-kc input is first used to produce outputs of 10 kc and 210 to 300 kc (in steps of 10 kc) for the oscilloscope test receptacles. A portion of the latter frequency is then mixed with the input from unit 2 and unit 5 to produce a final output from unit 6 of 10.4 to 10.5 mc (in steps of 10 cps), which is fed to the 100-kc step generator (unit 8).

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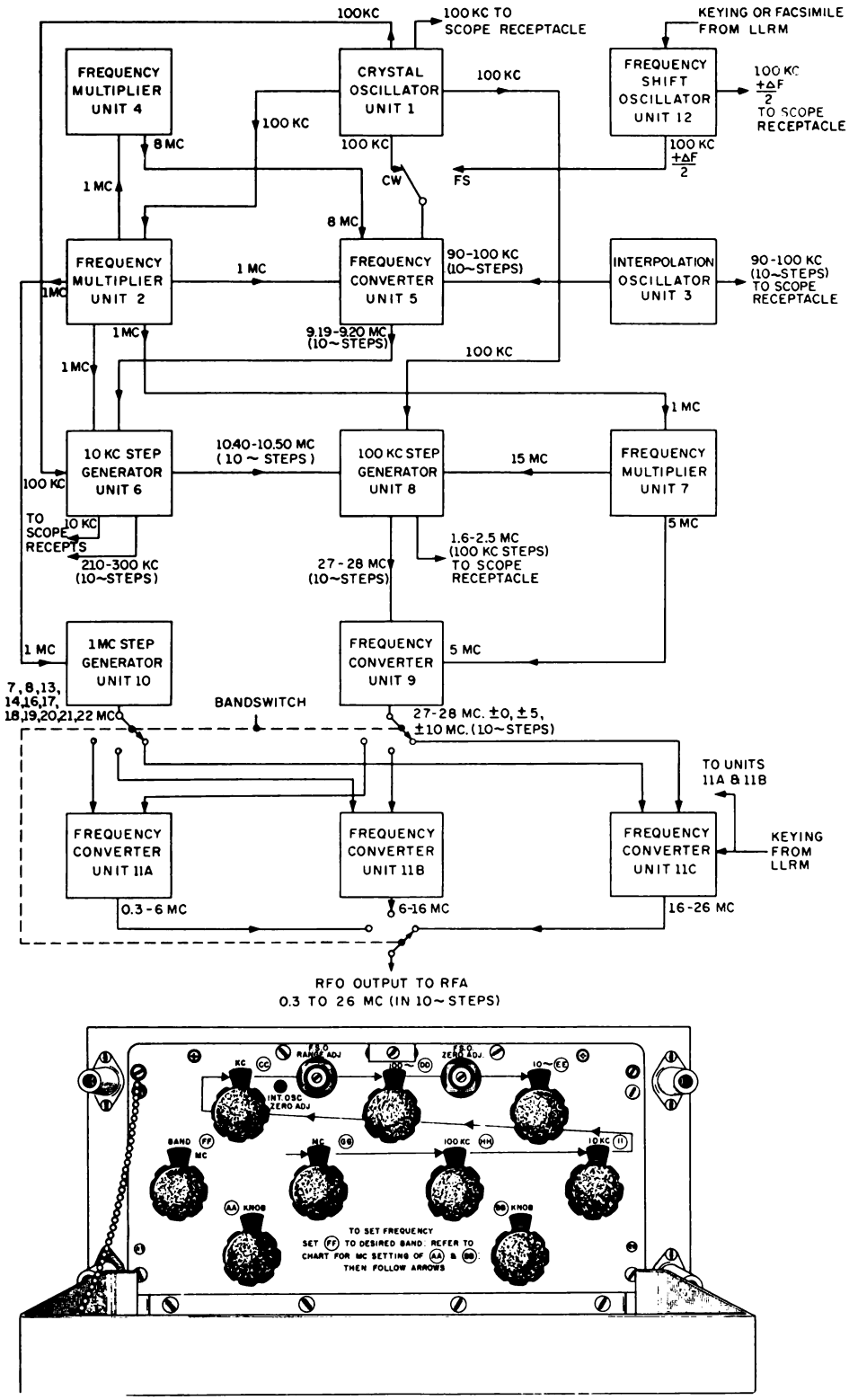
100KC
100 KC TO SCOPE RECEPTACLE
KEYING OR FACSIMILE FROM LLRM
FREQUENCY MULTIPLIER UNIT 4
CRYSTAL OSCILLATOR UNIT 1
FREQUENCY SHIFT OSCILLATOR UNIT 12
100 KC ±ΔF/2 TO SCOPE RECEPTACLE
8 MC
100 KC
100 KC ±ΔF/2
CW
FS
1 MC
FREQUENCY MULTIPLIER UNIT 2
FREQUENCY CONVERTER UNIT 5
INTERPOLATION OSCILLATOR UNIT 3
90-100 KC (10~STEPS)
90-100 KC (10~STEPS) TO SCOPE RECEPTACLE
9.19-9.20 MC (10~STEPS)
10 KC STEP GENERATOR UNIT 6
10.40-10.50 MC (10 ~ STEPS)
100 KC STEP GENERATOR UNIT 8
15 MC
FREQUENCY MULTIPLIER UNIT 7
5 MC
TO SCOPE RECEPTS
10 KC
210-300 KC (10~STEPS)
27-28 MC (10~STEPS)
1.6-2.5 MC (100 KC STEPS) TO SCOPE RECEPTACLE
1 MC STEP GENERATOR UNIT 10
FREQUENCY CONVERTER UNIT 9
7,8,13, 14,16,17, 18,19,20,21,22 MC
BANDSWITCH
27-28 MC. ±0, ±5, ±10 MC. (10~STEPS)
TO UNITS 11A & 11B
FREQUENCY CONVERTER UNIT 11A
FREQUENCY CONVERTER UNIT 11B
FREQUENCY CONVERTER UNIT 11C
KEYING FROM LLRM
0.3-6 MC
6-16 MC
16-26 MC
RFO OUTPUT TO RFA
0.3 TO 26 MC (IN 10~STEPS)
INT. OSC ZERO ADJ
TO SET FREQUENCY SET (FF) TO DESIRED BAND. REFER TO CHART FOR MC SETTING OF (AA) & (BB). THEN FOLLOW ARROWS

Figure 5-5. —Functional block diagram of radio-frequency oscillator and external view of control panel.

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Unit 7

The frequency multiplier (unit 7) receives a 1-mc signal from unit 2 and multiplies it to provide first a 5-mc signal for the frequency converter (unit 9) and, secondly, a 15-mc signal for the 100-kc step generator (unit 8). Unit 8 uses the 100-kc input from the crystal oscillator to produce an output of 1.6 to 2.5 mc, adjustable in 100-kc steps. A portion of this output may be fed to an oscilloscope. The remaining portion is mixed with the inputs from unit 6 and unit 7 to produce an output of 27 to 28 mc, which is adjustable in steps of 10 cps. This output is fed to the frequency converter (unit 9).

Unit 9

Unit 9 mixes the 27 to 28 mc input from unit 8 with the 5-mc input from unit 7 to produce frequencies of 27 to 28 mc (±0, ±5, or +10 mc), adjustable in 10-cps steps. These frequencies are amplified by the remainder of unit 9 and then fed to one wafer of the bandswitch. The one-mc step generator (unit 10) receives a 1-mc signal from unit 2 and passes it through a harmonic generator that produces frequencies of 7, 8, 13, 14, 16, 17, 18, 19, 20, 21, and 22 mc. These frequencies are amplified by the remainder of unit 10 and fed to the second wafer of the bandswitch.

Unit 11

The outputs of units 9 and 10 are fed to one of the three final converters (units 11A, 11B, and 11C), only one of which is used at a time. The output frequency desired determines which of the three converters will be used. The chosen unit is connected to the outputs of unit 9 and unit 10, and to the radio frequency oscillator mounting (unit 14) by the bandswitch. Unit 11A is used for frequencies between 0.3 and 6 mc, unit 11B for the 6- to 16-mc range, and unit 11C for outputs of 16 to 26 mc.

Unit 14

The mounting (unit 14) serves as a rack into which the other units of the radio frequency oscillator are connected. Unit 14, in turn, plugs into the transmitter group cabinet wiring. The control indicator, which mounts on the unit 14 front panel contains controls for antenna tuning and an indicator for measuring the standing wave ratio existing on the r-f transmission line.

The block diagram of figure 5-5, A, shows the signal distribution and signal progression through the RFO. In the discussion that follows, the circuits of the radio frequency oscillator are analyzed to show how the signal advances through each particular subunit. As you proceed through this discussion, it would be wise to make frequent reference to the block diagram of figure 5-5, A, particularly when moving from one subunit to another.

Unit 1 includes a crystal oscillator and four cathode followers. The crystal oscillator unit is one of the main frequency sources of the transmitter. Its 100-kc signal output is processed by other subunits of the radio frequency oscillator to obtain high frequencies. Other frequency sources in the radio frequency oscillator, such as the 10-kc step generator and the 100-kc step generator, are synchronized by the crystal oscillator and are thus locked on an accurate, crystal-controlled frequency standard. Two additional frequency sources, the interpolation oscillator and the frequency shift oscillator, are provided with frequency generators that are not locked to the crystal oscillator. However, the output of these frequency converters may be compared with the output of the crystal oscillator. The comparison is accomplished by means of a zero adjust indicating circuit located in unit 14, which employs an electron ray tube.

CRYSTAL OSCILLATOR (UNIT 1)

Crystal Oscillator

The crystal oscillator, V1 (fig. 5-6) is a modified electron-coupled Colpitts oscillator that operates at a frequency of 100 kc. It generates this frequency with an accuracy of 1.5 parts per million over the temperature range -4 to +122 degrees fahrenheit. Minor tuning adjustments may be made by means of the two slugs of the permeability tuned inductor, L1. One slug is factory adjusted; whereas the other slug may be adjusted in the field. The total range of the field adjustment is one cycle. L2 and C1 are provided to suppress parasitic oscillations. The output of V1 is a triangular waveshape with high harmonic content. This output is coupled through C2 to the control grid of the buffer cathode follower, V2A.

Cathode Followers

The cathode followers, V2 and V3, provide a method of coupling the output of the crystal

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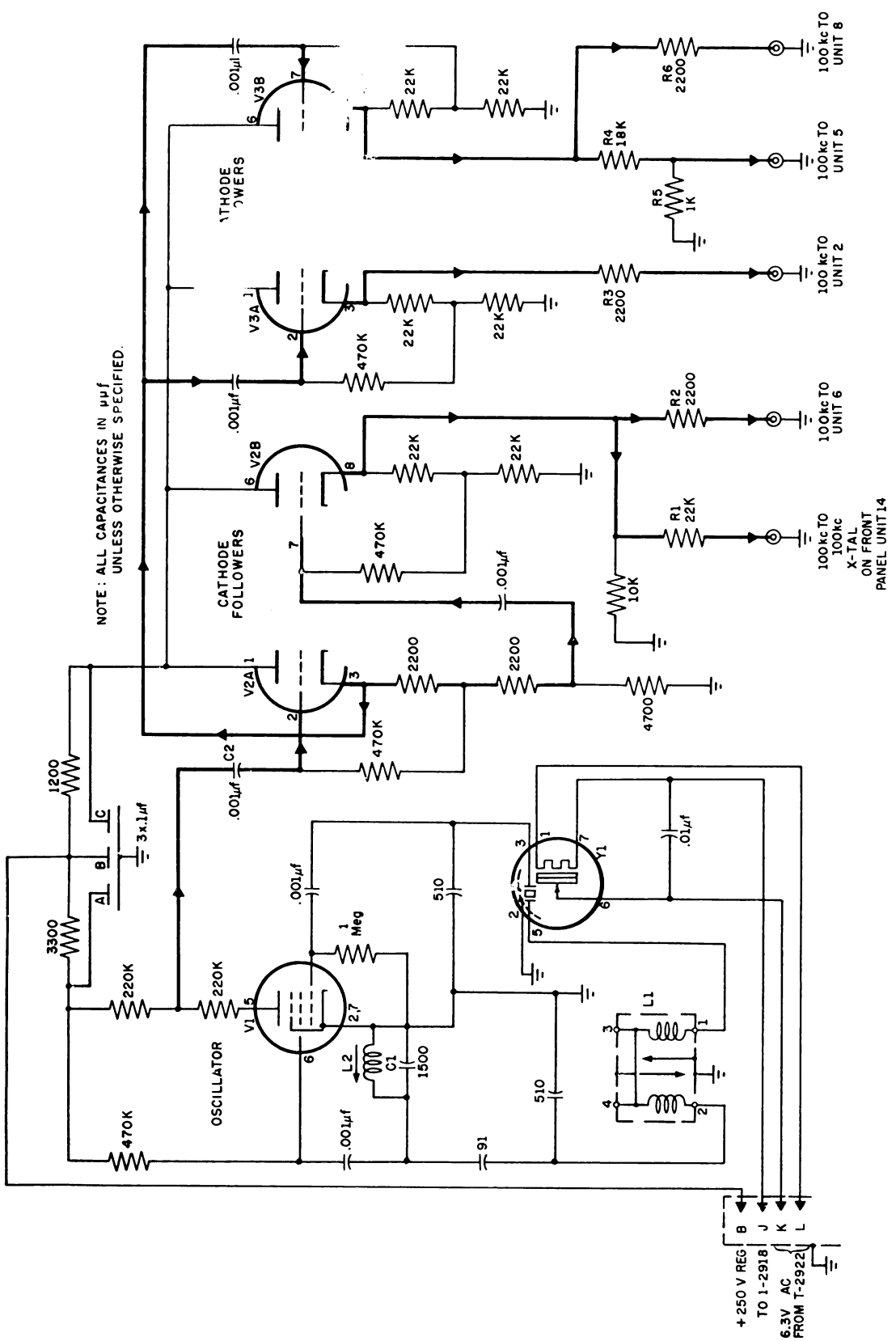

Figure 5-6. --Crystal oscillator, unit 1, schematic diagram.

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oscillator to other subunits of the radio frequency oscillator. They prevent loading of the crystal oscillator, thereby helping it to maintain its frequency stability. Cathode follower V2A receives the crystal oscillator output and distributes it to cathode followers V2B, V3A, and V3B. Thus, V2A serves as a buffer, or isolating stage, which isolates the oscillator from the output load. The 100-kc output of V2B is fed through isolating resistor R1 to the oscilloscope test receptacle in unit 14 and through isolating resistor R2 to unit 6. The 100-kc output of V3A is fed to unit 2 through isolating resistor R3. The corresponding frequency output of V3B is fed to voltage divider, R4 and R5, and to unit 8 through isolating resistor R6. The portion of the signal that is developed across R5 is fed to unit 5. In this case, R4 serves as an isolating resistor. The 1-mc signal with which the 100-kc signal is heterodyned in unit 5 is of relatively low amplitude and is applied to the suppressor grid, whereas the 100-kc signal is applied to the control grid of the mixer. The 100-kc output is therefore reduced in amplitude to match the low amplitude of the 1-mc input to unit 5, and to compensate for the higher amplification of the 100-kc signal resulting from its application to the mixer control grid.

FREQUENCY MULTIPLIER (UNIT 2)

The frequency multiplier (unit 2) receives the 100-kc crystal oscillator output of unit 1 via the cathode follower, V3A (fig. 5-7). Unit 2, by means of multipliers and harmonic filters, produces a 1-mc signal, which is delivered to units 4, 5, 6, 7, and 10. The multipliers of unit 2 consist of a frequency quintupler, V1, and a frequency doubler, V2.

Frequency Quintupler

The frequency quintupler, V1, receives from unit 1 the 100-kc output signal through J1, through transmission line terminating resistor R1, and through the grid leak network, R2 and C1. The quintupler (operated class C) delivers 100 kc and harmonics to filter Z1. Z1 is called a filter because its double-tuned tank circuits are capable of selecting the desired frequency. They are both tuned to 500 kc. Frequencies other than the fifth harmonic, 500 kc, are attenuated by this filter. Test point J2 is provided for checking the input signal amplitude, which should be approximately 13 volts.

Frequency Doubler

The frequency doubler, V2, also operates Class C, and receives the 500-kc output from Z1. The doubler stage delivers 500 kc and harmonic frequencies to a 1-mc filter comprising Z2 and Z3, capacitively coupled by C2.

Frequencies other than the second harmonic, 1 mc, are attenuated by these filters because the tuned circuits are resonant at 1 mc. The output of Z3 is impressed across the capacitive voltage divider, C3 and C4.

Cathode Followers

The cathode followers, V3 and V4, function to isolate the multipliers from the units to which the 1-mc signal is applied. The output of Z3 is applied to the cathode followers, V3 and V4, from the capacitance voltage divider, C3 and C4. Cathode follower output of V4B is fed to mixers, which require small input amplitudes. The outputs of V4B are therefore attenuated by the resistors, R3 and R4, respectively. The outputs of V3A, V3B, and V4A are fed to multipliers, and require no reduction in amplitude.

FREQUENCY MULTIPLIER (UNIT 4)

The frequency multiplier (unit 4) comprises a frequency quadrupler, V1, and a frequency doubler, V2 (fig. 5-8). Unit 4 receives the 1-mc output of unit 2, and by means of multipliers and harmonic filters produces an 8-mc signal. All capacitors represented by dashed lines are built in-into the tube socket and are .001 μf.

Frequency Quadrupler

The frequency quadrupler accepts the 1-mc output of unit 2 (previously described), which is developed at the control grid of V1 by R1. This stage (operated Class C) produces frequencies that are filtered by Z1 to attenuate all harmonics except the fourth, which is 4 mc (the resonant frequency of the tuned circuits). Test point J1 is provided to check the signal amplitude, which should be approximately 8.8 volts across the terminating resistor, R1. The 4-mc output of Z1 is fed to the grid of the doubler, V2.

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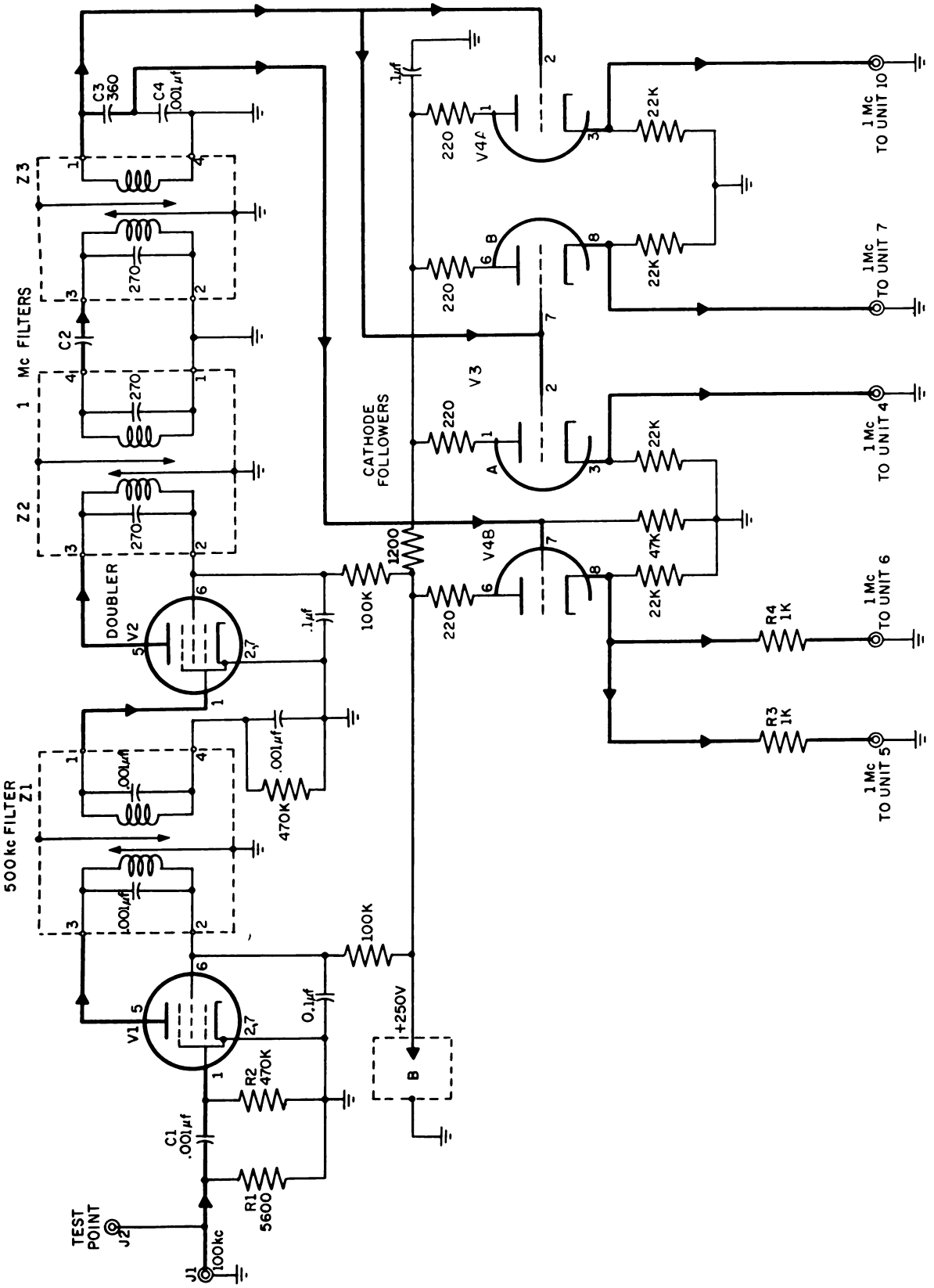

Figure 5-7. —Frequency multiplier, unit 2, schematic diagram.

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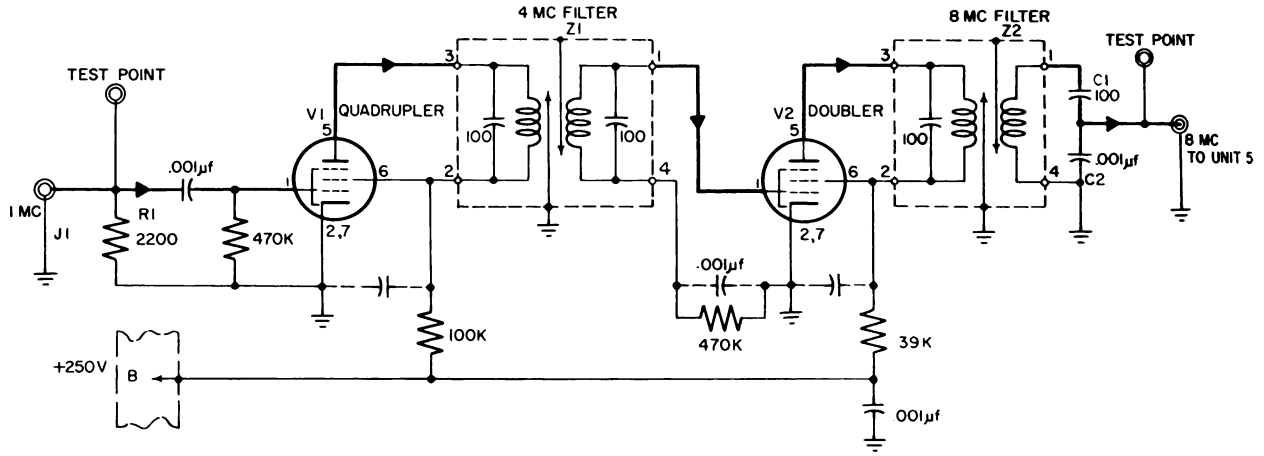

Figure 5-8. —Frequency multiplier, unit 4, schematic diagram.

Frequency Doubler

The frequency doubler stage delivers an 8-mc output to the filter, Z2, tuned to 8 mc, which attenuates all frequencies other than 8 mc. The final output is fed to unit 5 after being reduced by capacitance voltage divider, C1 and C2, to a level of approximately 1.2 volts.

INTERPOLATION OSCILLATOR (UNIT 3)

The interpolation oscillator (unit 3) generates frequencies in the range of 90 to 100 kc, which is made variable in steps of 10, 100, and 1000 cycles per second (fig. 5-9). To obtain maximum stability, the main frequency-controlling elements are housed in an oven at a constant 158 degrees fahrenheit. The output frequency is varied by means of decade switches that place inductors and capacitors of the required size in the oscillator main tank circuit.

Electron-Coupled Hartley Oscillator

An electron-coupled Hartley oscillator serves as the frequency source in unit 3. The main tank circuit of the oscillator, comprising C1 and L1 and associated elements, are located in the oven, a part of Z1. Trimmer coil L2 makes possible the adjustment of the resonant frequency of the tank circuit to exactly 100 kc. L2 and the oven assembly constitute Z1. The oscillator frequency may be changed by switching additional inductors and capacitors in or out of the main tank circuit. S1, S2, and S3 are provided as controls on the radio frequency oscillator mounting unit 14 for accomplishing frequency selection. For stability, V1 uses regulated filament and plate supplies. To reduce interaction with other stages via the power supply leads, a three section plate decoupling filter, comprising R1, R2, C1A, C2, and C3, is used.

Oscillator Switch

The oscillator switch, S1, chooses the oscillator tuning components. Inductors L3 or any one of the inductors, L4 to L12, and any one of the corresponding capacitors, C4 and C5 to C13, may be placed in the tank circuit by means of switch S1. In this way, the oscillator frequency may be changed in 1 kc steps. Let us consider the following example. With switches S1, S1A, and S2 in position 9 and switch S3 in position 10, the main tank inductor, L1, is series connected through contact 9 of S1, inductors L12 through L4, inductor L3, trimmer L2, contact 9 of S2, and contact 10 of S3 to ground. With S1A on contact 9, the main tank capacitor, C1, is connected in parallel with C4 (top capacitor) of the capacitor assembly, Z2. Under these conditions, the oscillator frequency is 100 kc.

If S1 is placed in position 8, L12 will be shorted out through contact 8 of the switch, thus reducing the inductance in the tank circuit. This normally would result in a higher oscillator frequency if it were not for the fact that, at the same time, the main tank capacitor, C1, is connected in parallel with capacitor C13 through S1A. The value of C13 is sufficiently large to more than compensate for the decrease in

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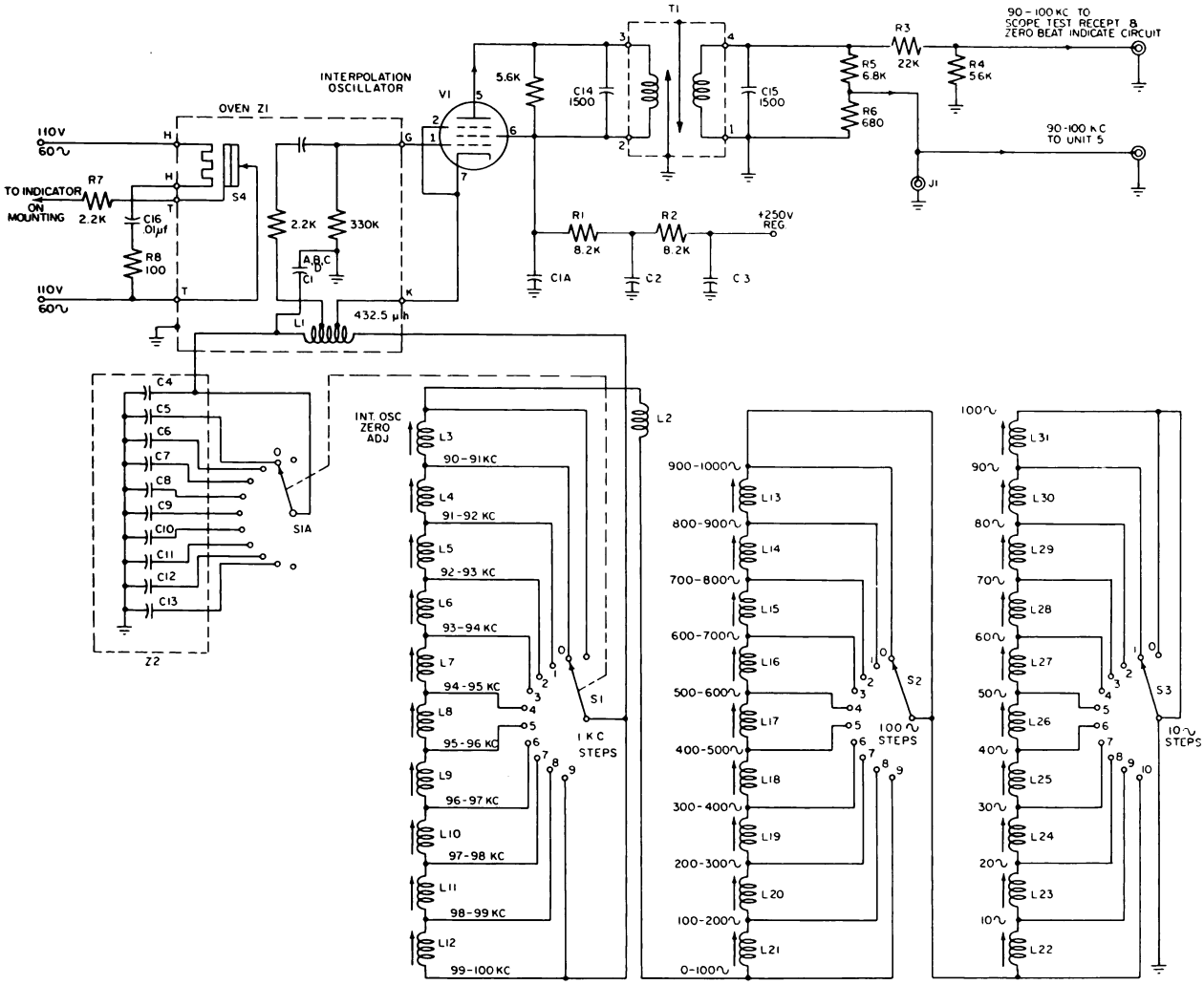

Figure 5-9. —Interpolation oscillator, unit 3, simplified schematic.

inductance. As a result of the increase in capacitance, the oscillator frequency is lowered to 99 kc. When S1 is placed in position 7, L12 and L11 are shorted out, and L1 is connected to an even higher capacitance than formerly, which more than compensates for the accompanying decrease in inductance. This action results in still a lower frequency of 98 kc. The process is repeated each time S1 is placed in a lower position, resulting in 1-kc step decreases of the oscillator frequency until 91 kc is reached. This occurs when S1 is in the 0 position. At this time, all of the inductors, L4 to L12, are shorted out, and the tank inductor, L1, is series connected to ground through L3 and trimmer L2 (S2 and S3 are in positions 9 and 10, respectively).

L3 is the INTERPOLATION OSCILLATOR ZERO ADJUST control. It permits adjustment of the oscillator output when this output is compared on the ZERO BEAT INDICATOR electron-ray tube with the crystal-controlled output of unit 1.

The method of changing the interpolation oscillator output frequency in 100 cps steps is similar to that just discussed for the 1-kc step switching. The 100-cps step switching is made possible through the action of switch S2.

With this switch in position 9, and S3 in position 10, the output of the 1-kc step switch, S1, is grounded by way of L2 through contact 9 of S2 and contact 10 of S3. Under these conditions, the 100-cycle switch, S2, has no effect on the oscillator frequency. However, when S2

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is placed in position 8, the inductors, L13 to L20 inclusive, are still shorted out, and the signal from the 1-kc step switch is fed through L21 to ground. The additional inductance introduced by L21 decreases the oscillator frequeny by 100 cycles. Similarly, when S2 is placed in position 7, inductor L20 is placed in the circuit, and the oscillator frequency is decreased by another 100 cycles. This process may be continued until S2 is placed in the 0 position (with S3 still in position 10) and a total decrease of 900 cycles is obtained in the oscillator frequency. Thus, S2 makes available frequency changes up to 900 cycles in 100-cycle steps.

When the 10-cycle step switch, S3, is placed in position 9, inductor L22 is added to the tank circuit. The additional inductance reduces the oscillator frequency by 10 cycles. The 10-cycle step decrease is continuous as switch S3 is moved from position 9 to its 0 position. When this position is reached, S3 will have made available oscillator frequency changes up to 100 cycles in 10 cycle steps.

The oscillator is provided with a tuned plate circuit output comprising the tunable transformer, T1, and capacitors C14 and C15. The resonant frequency of this circuit is 95 kc. R3, R4, R5, and R6 are separate voltage dividers from which the 90- to 100-kc output is obtained. In addition to providing the outputs, these resistors widen the bandpass of the tank circuit. The output taken from the junction of R3 and R4 is fed to the oscilloscope. The output taken from the junction of R5 and R6 is fed to a mixer of unit 5 (fig. 5-5). Test point J1 is provided for testing the output whose amplitude is approximately 0.3 volts.

Oven-Heater Circuit

The oven-heater circuit includes thermostat S4 and the heater element located inside the oven. R7, R8, and C16 are located outside the oven. R8 and C16 make up a spark suppressor circuit. When the heater is on, the interpolation heater on indicator lamp (not shown) is energized via the dropping resistor, R7.

FREQUENCY SHIFT OSCILLATOR (UNIT 12)

The frequency-shift oscillator (unit 12) which is shown in figure 5-10, makes possible frequency-shift keying and facsimile operation of the transmitter. The unit includes a 100-kc oscillator, which is capable of shifting its frequency ±1000 cps from 100 kc at a 240-cps rate during frequency shift operation. The oscillator is also capable of shifting up to 2000 cps, as determined by the amplitude of the photo input voltage, during facsimile operation.

Electron-Coupled Hartley Oscillator

The electron-coupled Hartley oscillator V1 is operated at a center frequency of 100 kc. In order to increase circuit stability the controlling tank circuit (comprising capacitors, C1, C2, C3, and tunable inductor L1, and associated elements) is contained in a temperature controlled oven. Plug-in inductor L1 is a coarse frequency control. Since the oscillator is the electron-coupled type, the tuned output tank (comprising T1, C4 and C5) filters the output and isolates the oscillator from the radio frequency oscillator subunits to which this output is connected. Load resistor R1 increases the bandpass of the output tank. R2 and R3 make up a voltage divider from which the output is coupled to unit 5 and to the oscilloscope test receptacle and zero beat indicator circuit.

Control Tube

The control tube, V2, is essentially a part of the oscillator circuit. A portion of the oscillator output is coupled to the grid of the control tube by C6. This tube functions to change the reactance of the tank circuit, and will therefore change the oscillator frequency. The frequency may be adjusted to 100 kc (in the absence of frequency shift or facsimile keying signals) by means of the coarse control, L1, and the fine control, R4. This control, called the Frequency Shift Oscillator (FSO) Zero Adjust is located on the control panel and provides a method of adjusting the cathode bias of the control tube. If the control tube grid voltage is made more positive, it will deliver a large capacitive reactive current to the tank. The increased leading current, which is equivalent to an increase amount of tank capacitance, causes the oscillator frequencies to decrease. If the grid voltage is made less positive, the oscillator frequency will increase.

Cathode Follower

The cathode follower output is connected through cathode (isolating) resistor R5 to the

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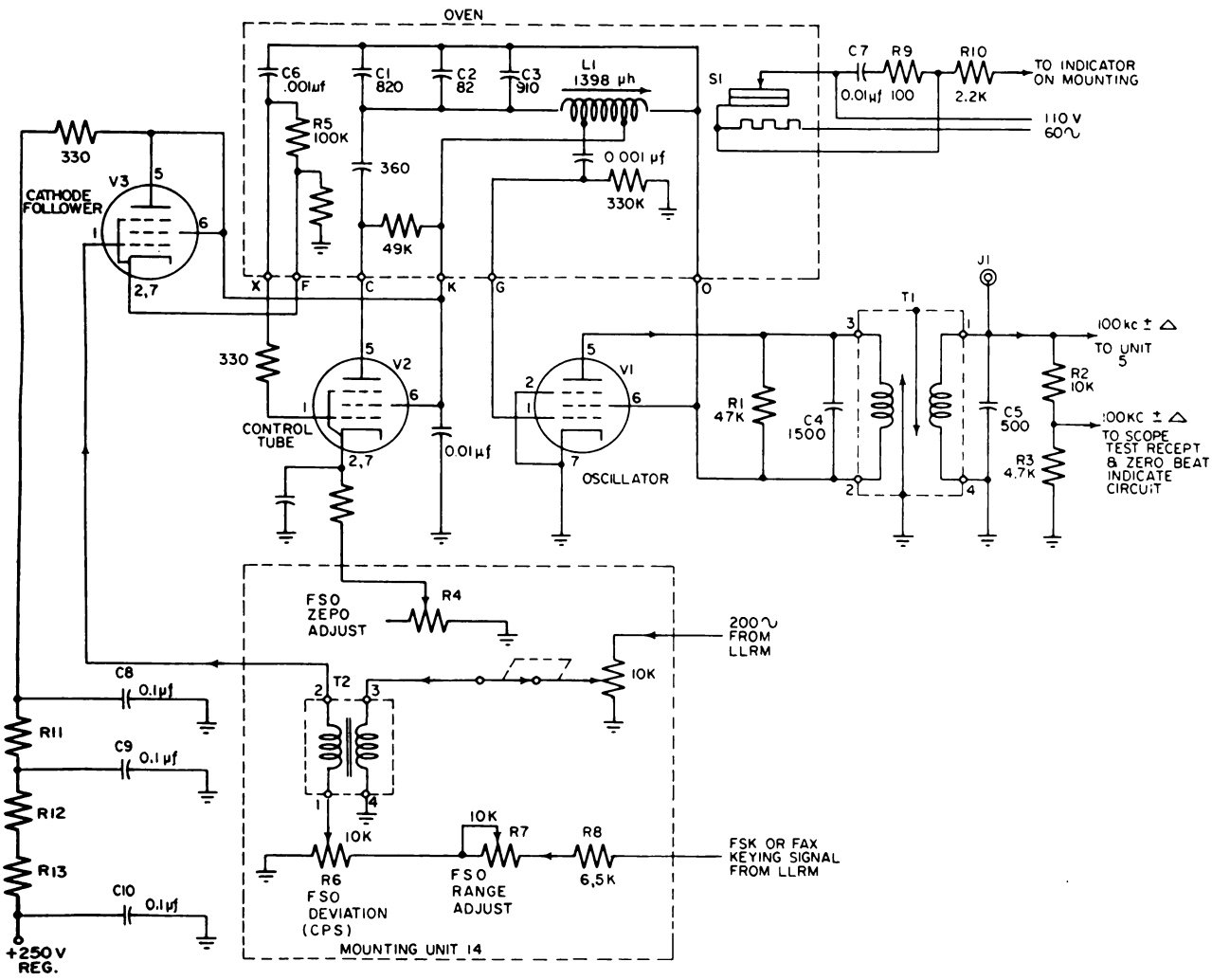

Figure 5-10. —Frequency shift oscillator, unit 12, simplified schematic.

grid of the control tube, V2. By this means, the oscillator frequency may be made to vary in accordance with the variations at the cathode follower grid. The grid of V3 is returned to ground through the secondary winding of T2 and a portion of potentiometer R6. The variable components are mounted on the front panel of the radio frequency oscillator. For frequency shift keying operation, the "frequency shift deviation" control may be set to provide an oscillator frequency deviation of up to ±1000 cycles as determined by the signal from the low level radio modulator.

Frequency-Shift Deviation Control

The frequency-shift deviation control, R6, is calibrated directly in cycles per second deviation from the nominal carrier frequency. When R6 is set to its midposition (500), a 1000-cps space-to-mark frequency shift, corresponding to a 500-cycle deviation on either side of the carrier, is provided. For facsimile operation, the equipment must be adjusted to provide a total frequency shift between 500 and 2000 cycles for a facsimile set output keying voltage of +20 volts d-c in normal operation.

The frequency-shift deviation control, R6, receives its operating potential from a voltage divider, which includes R7 and R8. R7 may be set so that the deviation dial reads correctly. The d-c potential developed between the arm of R6 and ground is applied to the control grid of the cathode follower by way of the 1-2 winding of T2. The FSK or FAX keying signals from

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the low-level radio modulator are rectangular pulses, which vary the grid-cathode bias of V3. At the same time a 200-cps signal from the low-level radio modulator may be used to phase modulate the FSK or FAX signal by way of the mutual coupling of T2.

The modulated keying pulses impressed on the cathode follower grid vary the reactive current in the control tube by varying its bias. The control tube, therefore, frequency modulates the oscillator in accordance with the amplitude of the keying pulses at the grid of the cathode follower. The cathode follower isolates the oscillator circuit, and T2 isolates the cathode follower (no signal d-c component) from the 200-cycle source in the low-level radio modulator.

Heater Circuit

The heater circuit of the frequency-shift oscillator is provided with a thermostat, S1. R9 and C7 constitute a spark suppressor circuit. R10 is a dropping resistor for the frequency-shift heater on indicator lamp.

The decoupling filter, comprising the capacitors, C8, C9, and C10, and resistors, R11, R12, and R13, is provided to minimize interstage coupling through the common power source. The power supply provides a regulated +250 volts to the plates and screen grids, and a regulated filament voltage for the tubes in the frequency-shift oscillator circuit.

Frequency-Shift Oscillator Output

The output of unit 12 (100 kc $\pm \frac{\Delta f}{2}$ where $\frac{\Delta f}{2}$ represents frequency deviation) is fed to unit 5 and to the oscilloscope test receptacle in unit 14. J1 is provided for checking the output signal amplitude, which should be approximately 12.3 volts.

The frequency-shift oscillator is used only when transmitting FSK or FAX signals. During c-w and phone operation, the crystal oscillator (unit 1) is used as the master frequency source instead of unit 12.

ELECTRONIC FREQUENCY CONVERTER (UNIT 5)

The frequency converter (unit 5) receives five frequencies from other subunits of the radio frequency oscillator (fig. 5-11). Four of these frequencies are constantly mixing to produce an output in the range of 9.19 to 9.2 mc in 10-cycle steps through a process that will be explained in the next paragraph. The frequencies received are 1 mc from unit 2, 100 kc from unit 1, 100 kc $\pm \Delta f$ (deviation from rest frequency caused by frequency shift signals) from unit 12, 90 to 100 kc from unit 3, and an 8-mc signal from unit 4. The 1000-kc input from unit 1 is used during c-w and phone operation only. During FSK and FAX operation, the 100 kc $\pm \frac{\Delta f}{2}$ or Δf signal from unit 12 is fed to the input of the frequency converter, replacing the 100 kc from unit 1.

First Mixer

The first mixer, V1, receives a 1-mc signal from unit 2, and an additional input signal of 100 kc from unit 1, or a 100 kc $\pm \frac{\Delta f}{2}$ from unit 12. The choice of the latter two input signals is dependent upon the mode of operation of the transmitter. When the low-level radio modulator is supplying frequency-shift or facsimile keyed signals to unit 5, relay K1 is energized from -24 volts through the GROUND FOR FREQUENCY SHIFT line shown on the schematic diagram of the LLRM (considered in chapter 7). With the relay in this position, the input (100 kc $+ \frac{\Delta f}{2}$ or 100 kc $+ \Delta f$) from unit L2 is developed across R2 and applied to the control grid of the mixer stage, V1, through resistor R1 and the relay contacts. During phone and c-w operation the ground from the LLRM is removed, and relay K1 is deenergized (in the position shown). As a result, the input signal from unit 12 is grounded through R1 and the 100 kc input from unit 1 is applied to the mixer control grid. The 1-mc input from unit 2 is applied to the mixer suppressor grid. The mixer output is passed through a dual filter arrangement consisting of Z1 and Z2. The mixing of the 1-mc input from unit 2 and the 100-kc input from unit 1 yields a 1.1 mc output, which is applied to the suppressor grid of the second mixer.

Second Mixer

The second mixer comprises tube V2 unit 3, the interpolation oscillator, supplies unit 5, the frequency converter, with a 90- to 100-kc signal, which is applied to the control grid of

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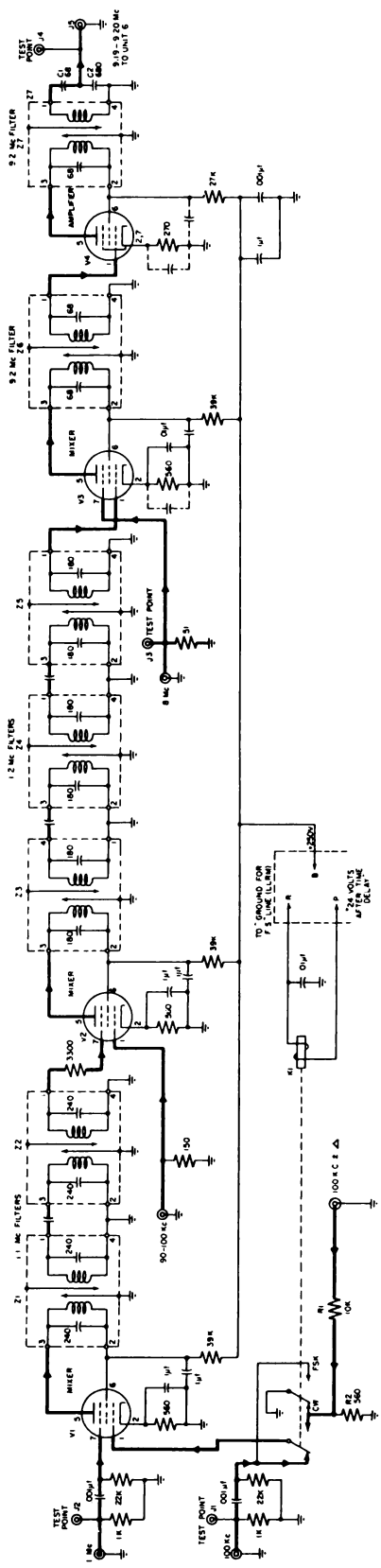

Figure 5-11. —Electronic frequency converter, unit 5, simplified schematic.

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the second mixer, V2. The conventional mixer output is filtered by sextuple-tuned filter arrangement comprising Z3, Z4, and Z5. Since the interpolation oscillator input to the mixer ranges over 10 kc (from 90 to 100 kc), the mixer output frequency to the control grid of the third mixer may be from 1.19 to 1.2 mc.

Third Mixer

The third mixer, V3, receives the 8-mc signal from unit 4 at its suppressor grid. The 9.19- to 9.2-mc output is filtered by Z6, amplified by V4, again filtered by Z7, reduced in amplitude by the voltage divider, C1 and C2, and fed to the output receptacle, J5. From this point, the output of unit 5 is transmitted to unit 6.

Test points J1, J2, J3, and J4 are provided to allow signal amplitude checks to be made at these points. The signal amplitudes should be as follows: at point J3, approximately 1.2 volts; at J2, approximately 1.2 volts: at J1, approximately 0.36 volts; and at J4, approximately 1.3 volts.

10-KC STEP GENERATOR (UNIT 6)

Unit 6 (fig. 5-12) produces an output of 10.4 to 10.5 mc, variable in 10 cps steps, and feeds this output to unit 8. It also generates a 10-kc output and an output of 210 to 300 kc, with the latter frequency group variable in steps of 10 kc. The unit contains a 10-kc step, phase locked oscillator comprising mixer, reactance tube, and oscillator. The output from these units is synchronized by the 100-kc frequency signal from the crystal oscillator of unit 1. To ensure the stability of the generated frequencies, the three tubes of the phase-locked oscillator, the blocking oscillator, and the shaper tube use a regulated filament supply. The output of the oscillator, V3 (210 to 300 kc) is mixed with a 1-mc signal, which arrives from unit 2 to the mixer, V5. The sum frequencies, ranging between 1.21 and 1.3 mc, is fed to mixer V6 and mixed with a 9.19- to 9.2 mc signal from unit 5. The sum frequencies, 10.4 to 10.5 mc, is fed to amplifiers V7 and V8, and then delivered to unit 8.

Isolating Amplifier

The isolating amplifier, V9, receives a 100-kc signal from unit 1 at its grid, coupled through C1. Amplifier V9 isolates the blocking oscillator, V4A, from the crystal oscillator (unit 1). In addition, the amplifier serves to provide sufficient output voltage to synchronize the blocking oscillator to 1/10 the crystal frequency.

Blocking Oscillator

The blocking oscillator, V4A, is employed in unit 6 to synchronize the step oscillator, V3. The block oscillator is a resistance-capacitance oscillator, whose free-running frequency depends primarily upon the R-C time constant in the control grid circuit. When tube current (in the blocking oscillator circuit) begins to flow through the plate winding of the transformer, T1, a voltage will be induced across the transformer, such that it will be positive toward the control grid of V4A. At first instant, the current through the tube is small, resulting in a small induced voltage across the transformer. The control grid voltage (at the time the tube begins to conduct) is zero. However, the positive going induced voltage, now being coupled across the transformer, is applied to the control grid of the blocking oscillator through the capacitor, C2. The tube, V4A, therefore, begins to conduct a heavier current. As the tube current increases, the amount of induced voltage likewise increases, and the control grid becomes still more positive. As the grid becomes more positive, the tube conducts a heavier current, and the process is continuous until the blocking oscillator tube conducts saturation current. During the time that the grid voltage is positive, C2 becomes charged as a result of grid current flow. As the tube reaches saturation, the plate current becomes constant, and the induced voltage across the transformer drops to zero. The capacitor, C2, immediately begins to discharge through R1, driving the grid far below the cutoff point. As capacitor C2 discharges, the grid voltage becomes less negative, and, in free-running operation, the grid potential is allowed to come just above the cutoff point. The cutoff time for V4A is therefore determined by the R-C time constant of the control grid network. If synchronizing forces are not applied at this time, the tube will again begin to conduct, and the process will be repeated.

In the blocking oscillator circuit of V4A, the grid recovery time is so chosen that only every tenth cycle from the isolating amplifier will effect the circuit operation. These cycles will reach the blocking oscillator at a time when the grid potential is sufficiently near the cutoff point to be raised above cutoff by the impressed signal.

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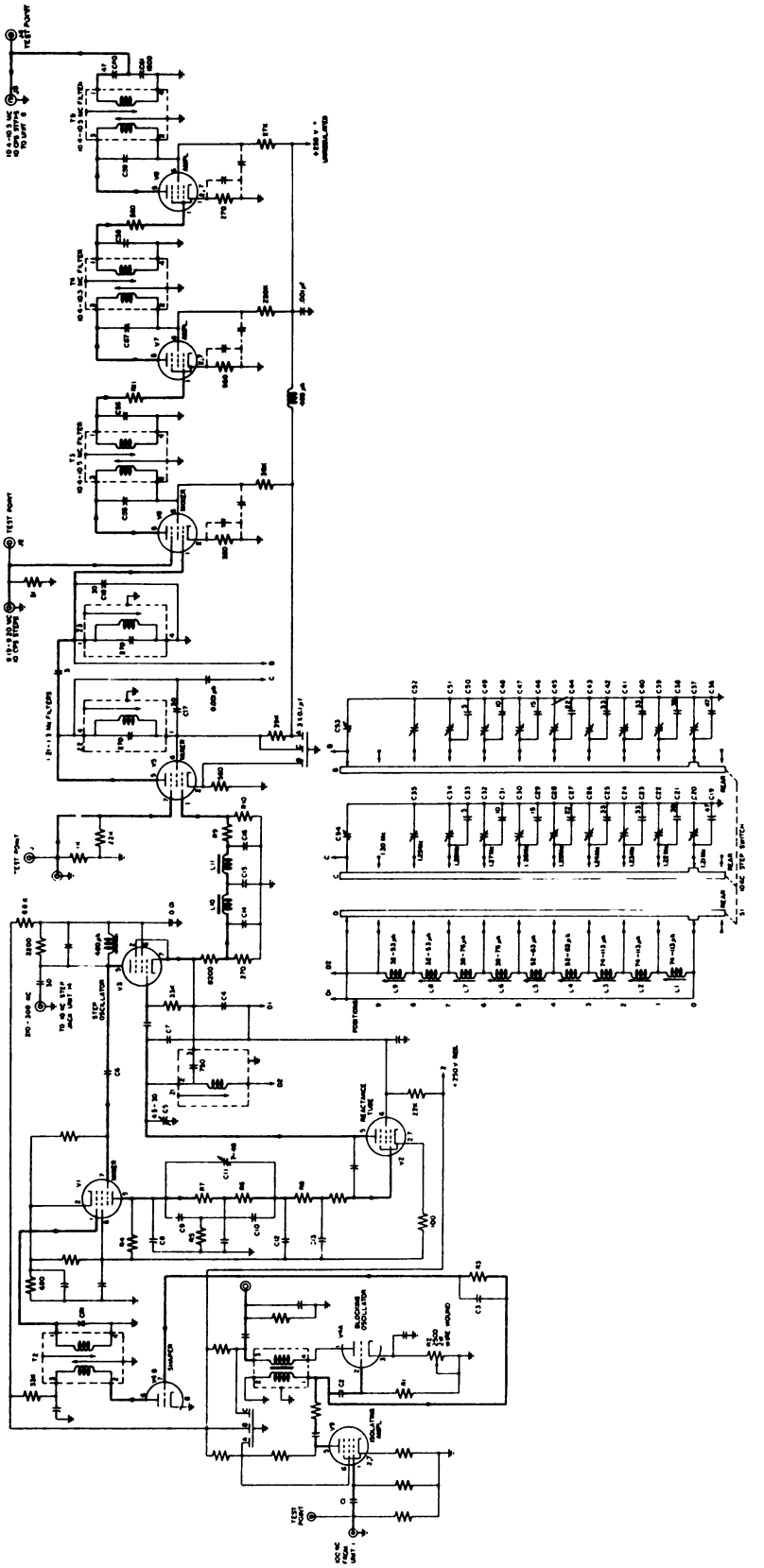

Figure 5-12. —Electronic frequency converter, unit 6.

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Therefore, the blocking oscillator frequency is synchronized to exactly 1/10 of the isolating amplifier output frequency, or 1/10 of the crystal oscillator frequency. The blocking oscillator natural frequency must be set at 10 kc or slightly lower to allow the isolating amplifier to synchronize its oscillations at a 10-kc rate. If it is set at slightly higher than 10 kc (that is, if the grid R-C time constant is decreased), the pulse from the isolating amplifier will arrive too late, and the oscillator will have operated prematurely. The free-running frequency of the oscillator (or the cutoff point of the blocking oscillator tube) may be varied by resistor, R2.

Shaper

The shaper tube, V4B, is provided to reduce the width of the blocking oscillator output pulse. As previously stated, the blocking oscillator is used to synchronize the step oscillator. To obtain more stable control of the step oscillator, the wide pulses from the blocking oscillator are passed through a peaking network C3 and R3. The pulse width is reduced to approximately 1 μs, and has an amplitude of 5 volts. The shaping circuit, V4B, with its damped tuned circuit, T2 and CR1, receives the pulse from the peaking circuit. The crystal, CR1, allows only the positive portion of the pulses to appear as the output of the shaper.

Step Oscillator

The step oscillator is locked-in to the 10-kc blocking oscillator by the mixer tube, V1, a filter network, and reactance tube V2. The modified Colpitts circuit of V3 makes up the 10-kc step oscillator. The tank circuit consists of Z1, C4, C7, and one or more of the inductors, L1 to L9, inclusive. The D section of S1 makes possible the selection of the inductors in accordance with the final output frequency desired from unit 6. The switch changes the output frequency in 10-kc steps from 210 to 300 kc. When S1 is in position "0", all the inductors, L1 through L9, are placed in the tank circuit, and the lowest oscillator frequency of 210 kc is obtained. When S1 is in position 1, L1 is shorted, and the inductance in the tank circuit is reduced. The oscillator frequency is therefore increased to 220 kc. Thus, the oscillator frequency may be increased (in steps of 10 kc) until in position 9, the output frequency is 300 kc. The frequency may be exacted to 300 kc by the use of the trimmer capacitor, C5. The output of 210 kc to 300 kc is fed to the 10-kc step oscilloscope test receptacle in unit 14.

Mixer

The mixer tube, V1, receives a portion of the 10-kc output of V3, which is coupled to the suppressor grid through capacitor C6. The non-sinusoidal output of the shaper tube comprising many harmonics is applied to the mixer control grid. The mixer output contains both d-c and a-c components. If the oscillator frequency is an exact multiple of 10 kc, the signal components reaching the grids of the mixer tube are the 10 kc and harmonics from the shaper tube, and the 210- to 300 kc signal from the step oscillator.

The d-c component of the mixer output is determined by the operating point of the tube, the amplitude of the input signals to the mixer grids, and the phase relationship existing between these two frequencies. The regulation of the power supplies is sufficient to hold the operating point of the tube constant, and to maintain the amplitude of the mixer output signals at a constant level. The phase relationship existing between the 10 kc output and step oscillator frequency is the most important factor in determining the d-c component of the mixer output.

Reactance Tube

The reactance tube, V2, develops the operating potentials for the mixer tube through a voltage divider in its cathode. The resistor, R4, is the mixer plate load. The operating d-c plate potential of the mixer tube is directly applied to the reactance tube control grid, and is sufficient to determine the reactance tube operating point, and therefore the magnitude of the reactive control of V2 in the oscillator circuit. The reactance tube causes the oscillator frequency to shift until the required reactance is contributed by V2 to maintain the desired frequency. If the oscillator frequency tends to increase, the phase relationship between the 10 kc and the oscillator frequency will change the mixer d-c output. The mixer change in potential is such that the magnitude of the capacitive reactance contributed by V2 will tend to decrease the oscillator frequency. In this manner, the 10-kc step oscillator is locked-in at the desired frequency.

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The composite a-c signal is not allowed to reach the control grid of the reactance tube, V2. This signal is filtered by a shunting capacitor, C8; a 10-kc rejection network (R5, R6, R7, C9, C10, C11); and a low-pass filter (C12, C13, and R8). The 10-kc filter is designed to ensure proper attenuation of the strong 10-kc signal from the mixer tube. The oscillator output is coupled to the 1.21- to 1.3 mc mixer, V5, by the low-pass filter network comprising L10, L11, C14, C15, and C16.

1.21- to 1.3 MC Mixer

The 1.21- to 1.3-mc mixer, V5, receives two signals; the oscillator output (210 to 300 kc), which is applied to the control grid, and a 1-mc signal from unit 2 applied to the suppressor grid. R9 and R10 form a voltage divider, which reduced the amplitude of the oscillator output signal. Test point J1 provides a place for making amplitude checks on the 1-mc signal, which should be approximately 1.2 volts. The mixer heterodynes the frequencies, and provides an output ranging from 1.21- to 1.3-mc. The mixer output circuit contains a double-tuned filter (consisting of Z2, Z3, C17, and C18) and the capacitors, which are switched in and out of the circuit by the sections C and B of the 10-kc step switch, S1. The D section of S1 containing the oscillator inductors and the B and C capacitor sections are ganged. It should be noted that S1 is the same step switch that selects the oscillator frequency in the range from 210 to 300 kc.

When step switch S1 is in position 1, capacitors C22, C21, C39, and C38 are used in the filter, and the filter is tuned to 1.22 mc. Likewise, the step oscillator at this time is tuned to 220 kc. As the switch position step numbers are increased, the capacitance in the filter is decreased by an amount sufficient to increase the resonant frequency by 10 kc for each step. Variable capacitors C20 through C35 and C37 through C52 are trimmers for adjusting the filter to the proper frequency for each position of S1. Variable capacitors C54 and C53 adjust the range of their respective capacitors and should only require resetting whenever the V5 is replaced.

10.4- to 10.5-MC Mixer

The 10.4- to 10.5-mc mixer, V6, receives the 1.21 to 1.3 mc output of V5 at its control grid and a 9.19- to 9.2-mc input from unit 5 at the suppressor grid. The heterodyning of these frequencies yields the 10.4- to 10.5-mc output, which is filtered by the double-tuned circuit comprising T3, C55, and C56. The output is coupled through R11 (a parasitic suppressor) to the control grid of the amplifier, V7. Approximately 1.3 volts should be measured at test point J2.

Amplifiers

The amplifiers, V7 and V8, amplify the 10.4 to 10.5-mc output of mixer V6. V7 filters its output through a double-tuned circuit comprising T4, C57, and C58. At the output of the second amplifier, the signal is again filtered by T5, C59, C60, and C61, and fed to unit 8 through J3. A signal amplitude of approximately 0.8 volts should be available at test point J4.

FREQUENCY MULTIPLIER (UNIT 7)

Unit 7 (fig. 5-13) receives a 1-mc signal from unit 2, multiplies it in a quintupler yielding a 5-mc output, and again multiplies the signal in a tripler circuit yielding a 15-mc output. The 5-mc signal is applied to unit 9, and the 15-mc output is coupled to unit 8.

Quintupler Stage

The quintupler stage, V1, receives a 1-mc signal from unit 2, which is coupled through capacitor C1 to its control grid. The signal is multiplied and passed through a filter, Z1, which attenuates all frequencies in the output except the fifth harmonic. The 5-mc signal is fed along dual paths to the control grids of the amplifier, V2, and through C2 to the tripler, V3. The 5-mc frequency is amplified by V2 and filtered by the double-tuned circuit of Z2. The center tap of capacitor voltage divider, C3 and C4, couples the 5-mc output to unit 9 via J1.

Tripler Stage

The tripler stage, V3, accepts the 5-mc signal, which is then multiplied and filtered by the double-tuned filter, Z3. The filter attenuates all frequencies except the third harmonic, 15 mc. The filtered output is amplified by V4 and filtered again by Z4. The final 15-mc output is coupled to unit 8. It is first reduced in amplitude by capacitor voltage divider C5 and C6

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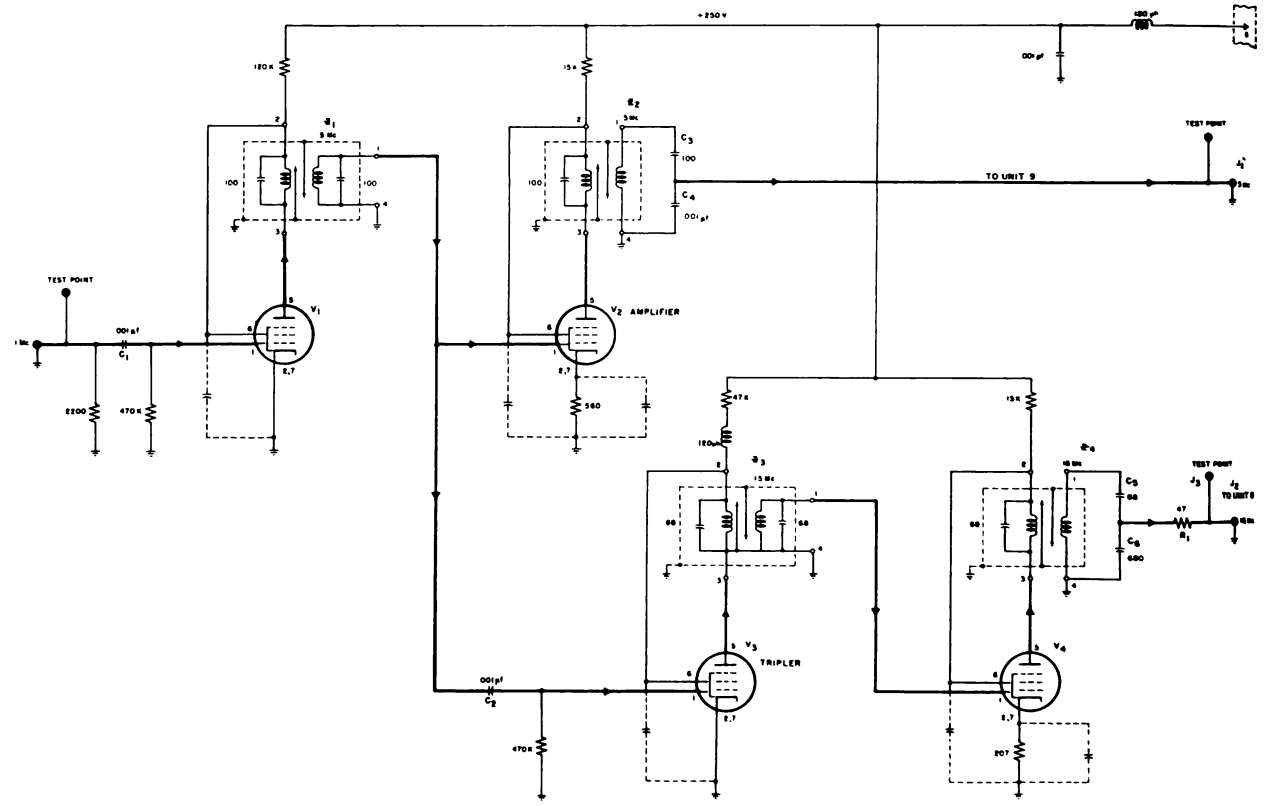

Figure 5-13. —Frequency multiplier, unit 7, simplified schematic.

and coupled by matching resistor R1 and J2. This output may be tested at J3 and is approximately 1.4 volts.

100-KC STEP GENERATOR (UNIT 8)

1.6- to 2.5-MC Phase-Locked Oscillator

The 1.6 to 2.5-mc phase-locked oscillator of unit 8 (fig. 5-14) may be tuned in steps of 100 kc. The circuit functions in a manner similar to that of the step oscillator of unit 6 already discussed. This oscillator is locked-in by the 100-kc, crystal-controlled oscillator of unit 1. The 1.6- to 2.5-mc output from the phase-locked oscillator is mixed with the 15-mc output of unit 7, resulting in frequencies ranging between 16.6 and 17.5 mc. These frequencies are then mixed with the 10.4- to 10.5-mc output of unit 6 to yield frequencies in the range of 27 to 28 mc. The 27- to 28-mc signal is made available in steps of 100 kc, 10 kc, and 10 cps. The unit contains amplifiers and filters and feeds its output to unit 9. To ensure better frequency stability, the heater voltage of the mixer tube, V1, the reactance tube, V2, and the oscillator tube, V3, is received from a regulated filament supply.

Clipper and Shaper

The clipper and shaper circuits contain components, which form the output pulse required to assure best control of the 1.6- to 2.5-mc phase-locked oscillator.

The 100-kc signal from unit 1 is received at J1 and coupled through C1 to the control grid of the clipper tube, V4A. The signal amplitude at J2 should be approximately 8.1 volts. The positive portion of each cycle of the 100-kc input is sufficient to allow the control grid to draw a charging current to C1. During the negative half of each cycle, the discharging capacitor and the negative going input are sufficient to cut off V4A. In this manner, both the positive and negative half of the 100-kc signal is clipped.

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The output of V4A has a square-wave appearance and is coupled through C2 to the control grid of the shaper tube, V4B. The output of the shaper is coupled to the transformer, T1, where it is further shaped by the oscillatory action of the transformer into a narrower pulse. This sharper pulse created by the high resonant frequency of the transformer will ensure more accurate control of the 1.6- to 2.5-mc phase-locked oscillator. Rectifier CR1 conducts during the negative portion of the cycle preventing any negative overshoot. The pulse is coupled by the secondary of the transformer to the control grid of the mixer tube, V1.

The oscillator, V3, a modified Colpitts, operates in the frequency range from 1.6 to 2.5 mc. Section G of the switch, S1, is controlled by a 100-kc frequency selection knob. The switch allows the inductors, L1 to L9, to be placed in the tank circuit of the oscillator, making possible selection of frequencies in 100 kc steps over the entire oscillator range. The oscillator frequency is further controlled by the reactance tube in a manner similar to that of the 210- to 300-kc oscillator of unit 6 (fig. 5-12). Mixer V1 heterodynes the 100-kc pulses from the shaper tube with the output of oscillator V3. Any existing phase difference in the two signals will affect the reactance tube, V2, such as to correct the oscillator for frequency drift. The rejection filter in the plate circuit of the mixer, in this case, is designed to suppress 100 kc, isolating the a-c component of the mixer plate voltage from the control grid of the reactance tube. The oscillator output is coupled to the control grid of the mixer, V5, by a low-pass filter consisting of L10, L11, C3, C4, and C5.

First Mixer

The oscillator output, 1.6 to 2.5 mc, is applied to the control grid, while the 15-mc signal from unit 7 is applied to the suppressor grid. The 15-mc input is received at J3 at an amplitude of approximately 1.3 volts and may be tested at J4. The mixer output, 16.6 to 17.5 mc, is filtered by the double-tuned circuit comprising L12, C6, C7, L13, C8, and C9, and the capacitors selected by the sections E and F of S1. The function of the switch is similar to that used in the oscillator circuit of unit 6, already discussed.

Second Mixer

The second mixer, V6, receives the output of V5 at the suppressor grid. The control grid receives the 10.4- to 10.5-mc input from unit 6 through J5. The magnitude of this input from unit 6 (fig. 5-12) should be approximately 0.8 volts, and may be checked at test point J6. The mixer output is filtered by the double-tuned circuit of Z1 and applied to the control grid of the amplifier, V7.

The 27- and 28-mc signal is amplified by V7 and filtered by the double-tuned circuit, Z2. The resistors, R1 and R2, act as loading resistors for the filter circuit. The signal is again amplified by V8 and filtered by Z3. The final 27- to 28-mc output signal is reduced by the capacitor voltage divider, C10 and C11, and fed to unit 9 through J7. The output should be approximately 1.2 volts.

FREQUENCY CONVERTER (UNIT 9)

Unit 9 (fig. 5-15) receives a 27- to 28-mc input, variable in steps of 10 cps, from unit 8. In addition, unit 9 receives a 5-mc signal from unit 7. After heterodyning in unit 9, the following frequencies may be obtained at the mixer output: 27 to 28 mc, 32 to 33 mc, 22 to 23 mc, and 37 to 38 mc. These frequencies correspond to the original, the sum, the difference, and the 5-mc second harmonic sum, respectively.

The mixer output is filtered, amplified, and sent to either unit 11A, 11B, or 11C, depending upon the desired final output frequency of the radio frequency oscillator.

Mixer

Mixer V1 receives the 27- to 28-mc output of unit 8 at its suppressor grid through J1. The 5-mc output of unit 7 is applied to the control grid through J1 and the front of wafer B of the 5-mc step switch, S1. The 5-mc signal is adjusted to the proper level for efficient mixing by R1 and R2 in the circuit of wafer B of S1. When an output between 27 and 28 mc is chosen, the mixer control grid is shorted to ground, and the 27- to 28-mc signal is passed to the output of the mixer tube. The components comprising the C section of S1 may be placed in parallel with the resistor, R3. The components of S2c may then be placed in parallel with those selected by S1c. Each filter made up by

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wafer C of the 5-mc step switch, S1, has a bandpass of 150 kc (±50 kc). In addition, the filters are designed to accept either the sum frequency, the difference frequency, the original frequency, or the 5-mc second harmonic frequency. The switch, S1, in position 0 allows an output from unit 9 of 22 to 23 mc; position 1, 27 to 28 mc; position 2, 32 to 33 mc; and in position 3, an output ranging from 37 to 38 mc. The operation of the switch is as follows: with S1 in position 2, L3 and C1 form a filter whose bandpass midpoint may be varied in 100-kc steps from 32.9 to 32.0 mc by the C section of S2.

If a final output of 15 mc is desired from the radio frequency oscillator, the switch S1 is placed in position 3. The output of unit 9 will be 37 to 38 mc. Unit 11B (discussed later in this chapter) will then mix the 37- to 38-mc signal with the proper frequency from unit 10 to yield the desired frequency of 15 mc. The position of the 100-kc step switch, S2, is controlled by the 100-kc knob located on the front panel and is always in the same position as the 100-kc step switch, S1, in unit 8, which is controlled by the same knob.

Amplifiers

Amplifiers V2, V3, and V4 increase the input signal strength necessary to provide mixing in the final frequency output stages of the radio frequency oscillator. This process is considered in the discussion of units 11A, 11B, and 11C.

The mixer output is applied to the control grid of the amplifier, V2. This stage amplifies and filters the signal. The filter circuits are similar to the mixer filter circuits already discussed, and their components are selected in a similar manner by section D of the 5-mc step switch, S1, and of the 100-kc step switch, S2. The filter circuits of V3 and their components are selected by section E of switches S1 and S2. To obtain approximately equal output for different frequencies, the screen and plate voltage of V2 and V3 are changed at different frequencies, thus varying the gain of the amplifiers. This is done by means of dropping resistors R4 through R7 of S1B and R8 through R16 located on the rear of wafer S2B. The tubes, V2 and V3, operate as conventional amplifiers.

The amplifier, V4, functions in a similar manner to V2 and V3. The process of changing the screen and plate resistor is not used in this stage.

The final output of unit 9, 27 to 28 mc ±0 ±5 + 10 mc, is applied through J3 to unit 11A, 11B, or 11C, depending on the desired output frequency of the radio frequency oscillator. The bandswitch, which determines the RFO output frequency, is located in unit 14. Test point J4 is provided for testing the output of unit 9 and should be approximately 1.6 volts. The B supply of unit 9 is decoupled by a filter network comprising L5, C2, C3, C4, R17, and R18.

1 MC STEP GENERATOR (UNIT 10)

Unit 10 (fig. 5-16) receives the 1-mc output of unit 2 and multiplies this input to 7, 8, 13, 14, 16, 17, 18, 19, 20, 21, or 22 mc. It employs a harmonic generator and a system of amplifiers with tuned filters in their plate circuits.

Harmonic Generator

The harmonic generator receives the 1-mc output of unit 2 at J1, which is applied across R1 to the control grid. The stage is operated class C, and, because of its low plate and screen voltage, limiting of the input signal is obtained. The harmonic generator distorts the input greatly, and, because of this, its output is rich in harmonics. J2 is provided for testing the amplitude of the input to V1, which should be approximately 8.6 volts.

Amplifiers

Amplifiers V2, V3, and V4 are similar to the harmonic generator. The output of V1, rich in harmonics, is applied to the control grid of the amplifier, V2, through C1. The latter stage distorts and amplifies the harmonic generator output. The amplifier output is very rich in harmonics and, therefore, very suitable for use in deriving the frequencies desired in the final output of unit 10.

The grid circuit of V3 consists of a tuned filter comprising C2 and any one of the inductors, L1 to L11, inclusive, chosen by section F of the 1-mc step switch, S1. This switch may be set in any one of 11 positions numbered from 0 to 10. In position 0, inductor L1 is selected. Its value is such that the plate filter circuit will be tuned to the seventh harmonic of 1 mc, and the plate circuit of V2 will attenuate all frequencies other than 7 mc. With S1 in position 1, all frequencies other than 8 mc will be attenuated, etc.

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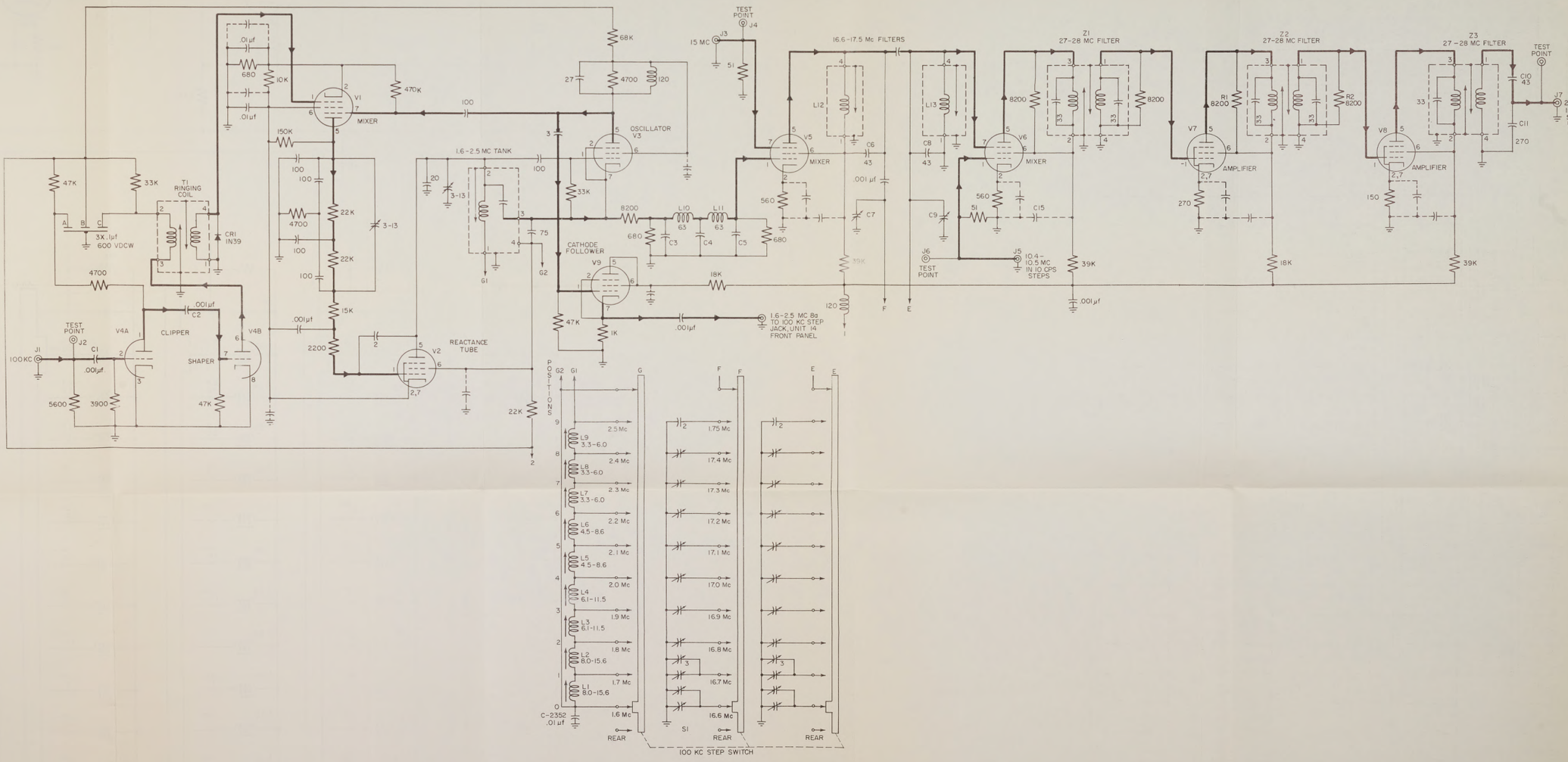

Figure 5-14. -100-kc step generator, unit 8, simplified schematic.

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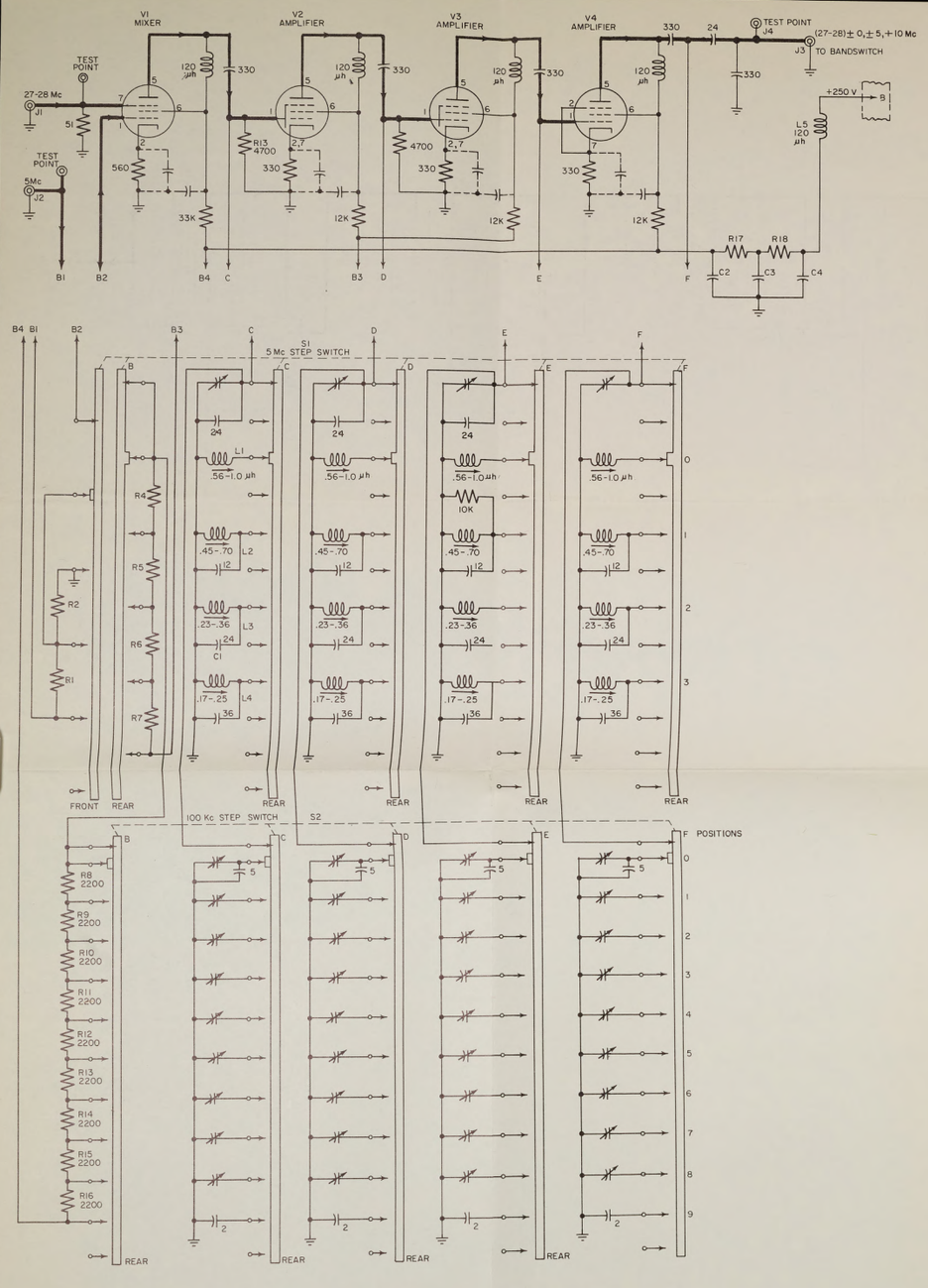

Figure 5-15. –Frequency converter, unit 9, simplified schematic.

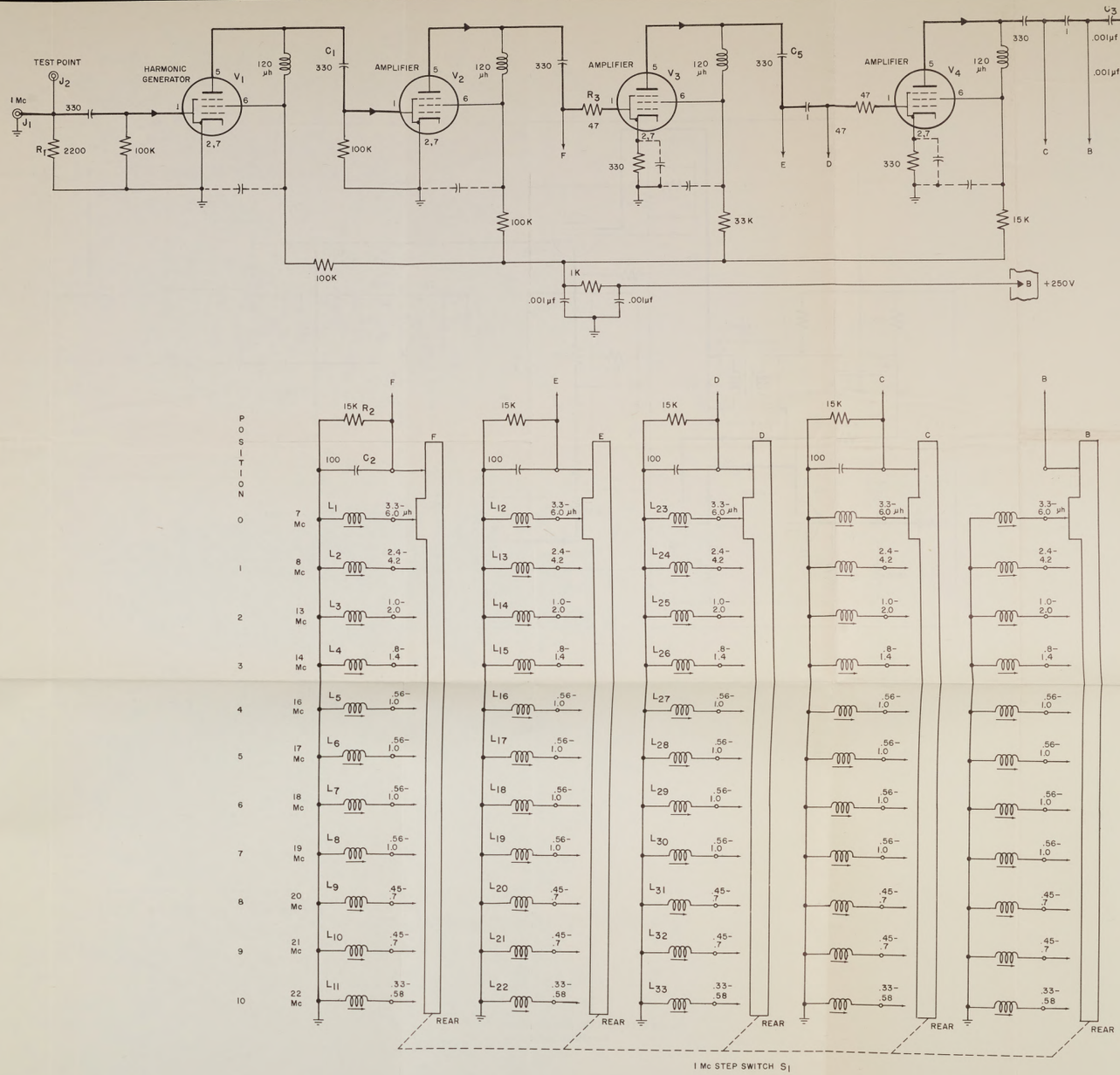

Figure 5-16. –1-mc step generator, unit 10, simplified schematic.

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To prevent the next high gain amplifier, V3, rom oscillating, load resistor R2 is provided o lower the Q in the filter circuit. In addition, R3 is provided in series with the control grid o further suppress oscillations, which might ccur.

The output of V3 is coupled to V4 by C5. Amplifier V4 is provided with a double-tuned nput filter circuit in which inductors L12 hrough L22 are selected by section E of S1, nd inductors L23 through L33 are selected by ection D of S1. The sections of S1 are ganged o that switching takes place simultaneously n all of the amplifier circuits. Amplifier V4 as, in addition, a double-tuned output filter ircuit. The frequency of the input filter is ontrolled by sections D and E of S1, and the requency of the output filter is controlled by ections C and B of S1. Table 5-1 shows the ositions of the 1-mc step switch, S1, and the utput frequency of unit 10 corresponding to ach position.

Table 5-1.--Positions of 1-MC Step Switch.

0	7
1	8
2	13
3	14
4	16
5	17
6	18
7	19
8	20
9	21
10	22

The final output of unit 10 is reduced in mplitude by the divider, C3 and C4. The out- ut is then applied to unit 11A, 11B, or 11C by he bandswitch, in unit 14, through coupling esistor R4 and receptacle J3. Test point J4 is rovided for testing the output of unit 10, which hould be approximately 0.4 volts. Unit 14 is a ack into which all of the subunits of the RFO re contained. Also contained in unit 14 are everal control circuits, such as the manual requency selection control circuits; the band- witch; the ovens for oscillator circuits, already iscussed; circuit indicating lamps; a door witch; the frequency-shift zero adjust control; he frequency-shift oscillator input circuits; he controls for units 3 and 12 adjustments; est receptacles for the oscilloscope; wires nterconnecting the RFO units; and the antenna ming control circuits.

The bandswitch located in unit 14 determines to which frequency converter unit (11A, 11B, or 11C) the output signals from units 9 and 10 are delivered, and from which units the final output of the RFO will be obtained.

FREQUENCY CONVERTER (UNIT 11A)

Units 11A, 11B, and 11C are the final frequency converters in the radio frequency oscillator. When the transmitter is set to operate on a frequency in the range of 0.3 to 6 mc, the bandswitch in unit 14 selects unit 11A (fig. 5-17) as the recipient of the output from units 9 and 10. Unit 11A then mixes the frequencies received from units 9 and 10 and delivers a frequency in the range of 0.3 to 6 mc to the radio frequency amplifier. Similarly, when the transmitter is set to operate on a frequency in the range of 6 to 16 mc, unit 11B is selected to receive the outputs of units 9 and 10. A frequency in the range from 16 to 26 mc will cause unit 11C to be selected as the final frequency converter. Unit 11A consists of a mixer, two amplifiers, filters, and a cathode follower output stage.

Mixer

The mixer, V1 (fig. 5-17) receives the output of unit 10, which is applied to the control grid through the capacitor, C1. The output of unit 9 is applied to the suppressor grid. Test points J1 and J2 are provided for testing the input amplitudes from the two units, which should be approximately 1.4 volts from unit 9 and 0.3 volts from unit 10. Mixer V1 produces an output in accordance with the two frequencies applied at its input. For example, if the desired output is from zero to one megacycle, the mixer will receive 22 to 23 mc from unit 9, and 22 mc from unit 10, and present the difference frequency at its output. A low-pass filter circuit comprising L1, L2, and L3, and C2 and C3 allows the passage of frequencies up to and including 6 mc. The output of the keying circuit of the low-level radio modulator (discussed in chapter 6) is applied to the control grid of the mixer, V1, through a decoupling filter comprising C4, R1, and grid resistor R2.

In frequency-shift keying and facsimile operation, a permanent ground is applied to V1, allowing the tube to conduct in response to the two signals from units 9 and 10. In hand-key and machine key operation, the keying signal,

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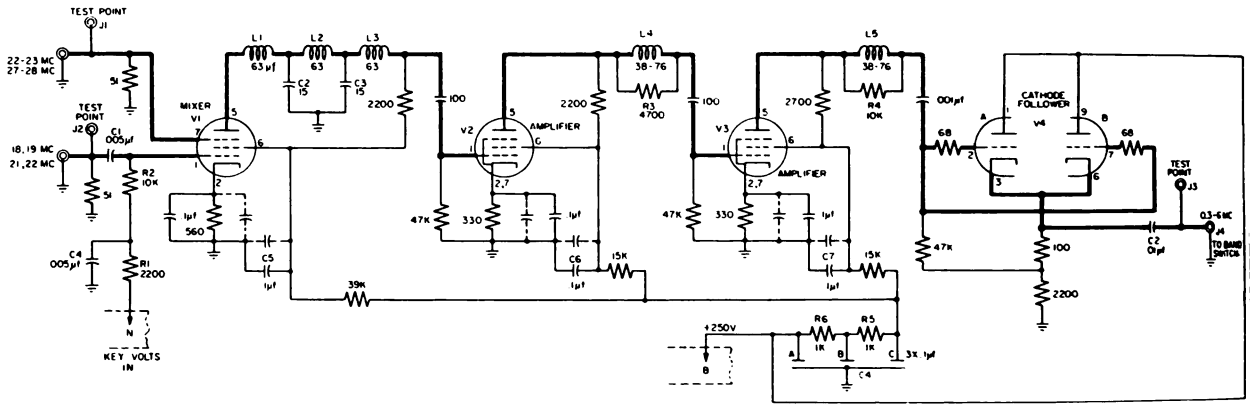

Figure 5-17.—Frequency converter, unit 11A, simplified schematic.

zero for a mark, and -35 volts for a space is applied to the control grid. During a mark the zero potential allows V1 to conduct while the -35 volts corresponding to a space cuts off the tube. The output from the radio frequency oscillator, during a space, is therefore reduced to zero. In phone operation, -35 volts is applied to the grid of V1 when the press-to-talk button is not depressed, cutting off the radio frequency oscillator in the same manner as the space signal. This method of keying is employed to ensure that the key-up radiation of the transmitter remains below the maximum allowable level of 400 μμv.

Amplifiers

Amplifiers V2 and V3 increase the signal amplitude necessary to provide sufficient driving input to the radio frequency amplifiers. The peaking coils, L4 and L5, are provided for video compensation to extend the high-frequency limit of the two amplifiers. The resistors, R3 and R4, in parallel with the coils limit the gain at high frequencies. A decoupling filter comprising C4, R5, and R6 is provided and serves all the stages of unit 11A. Further decoupling is accomplished in the plate circuits of the mixer and amplifiers by C5, C6, and C7. The two conventional and identical amplifiers amplify the mixer output and pass it to the cathode follower.

Cathode Follower

The cathode follower, V4, provides sufficient output amplitude (approximately 2 volts) to operate the radio frequency amplifier. The two triode sections are connected in parallel to provide the high-power output required from the cathode follower. V4 works into an output impedance of approximately 50 ohms, feeding its output through C8 to J3, J4, the bandswitch (not shown), and finally to the radio frequency amplifier.

FREQUENCY CONVERTER (UNIT 11B)

Unit 11B (fig. 5-18) receives the 13-, 14-, 16-, 17-, 18-, 19-, 20-, and 22-mc input from unit 10. From unit 9, unit 11B receives either a 22 to 23-, 27 to 28-, 32 to 33-, or 37 to 38-mc signal. The signals are mixed by V1 and amplified by V2, V3, and V4 (fig. 5-19). A 1-mc step switch, S1, and a 100-kc step switch, S2, are provided to select the proper filters for each stage required to produce the desired frequency and to reject others.

Table 5-2 shows the positions of the 1-mc step switch, S1, and the filter frequency range output of unit 11B corresponding to each position of the switch.

Table 5-2.--Positions of the GG Control.

6	6-7
7	7-8
8	8-9
9	9-10
0	10-11
1	11-12
2	12-13
3	13-14
4	14-15
5	15-16

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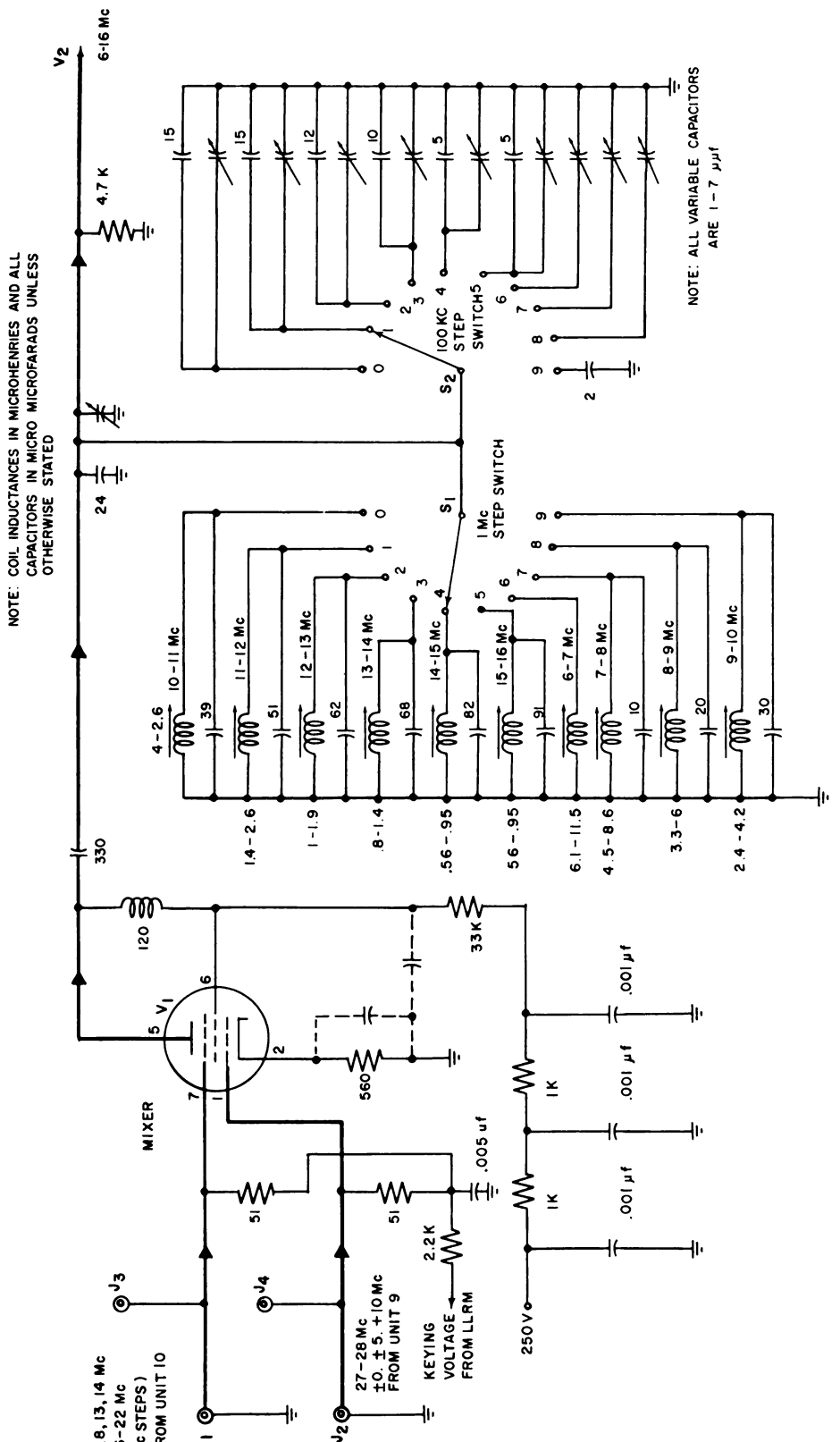

Figure 5-18.—Mixer, unit 11B, simplified schematic.

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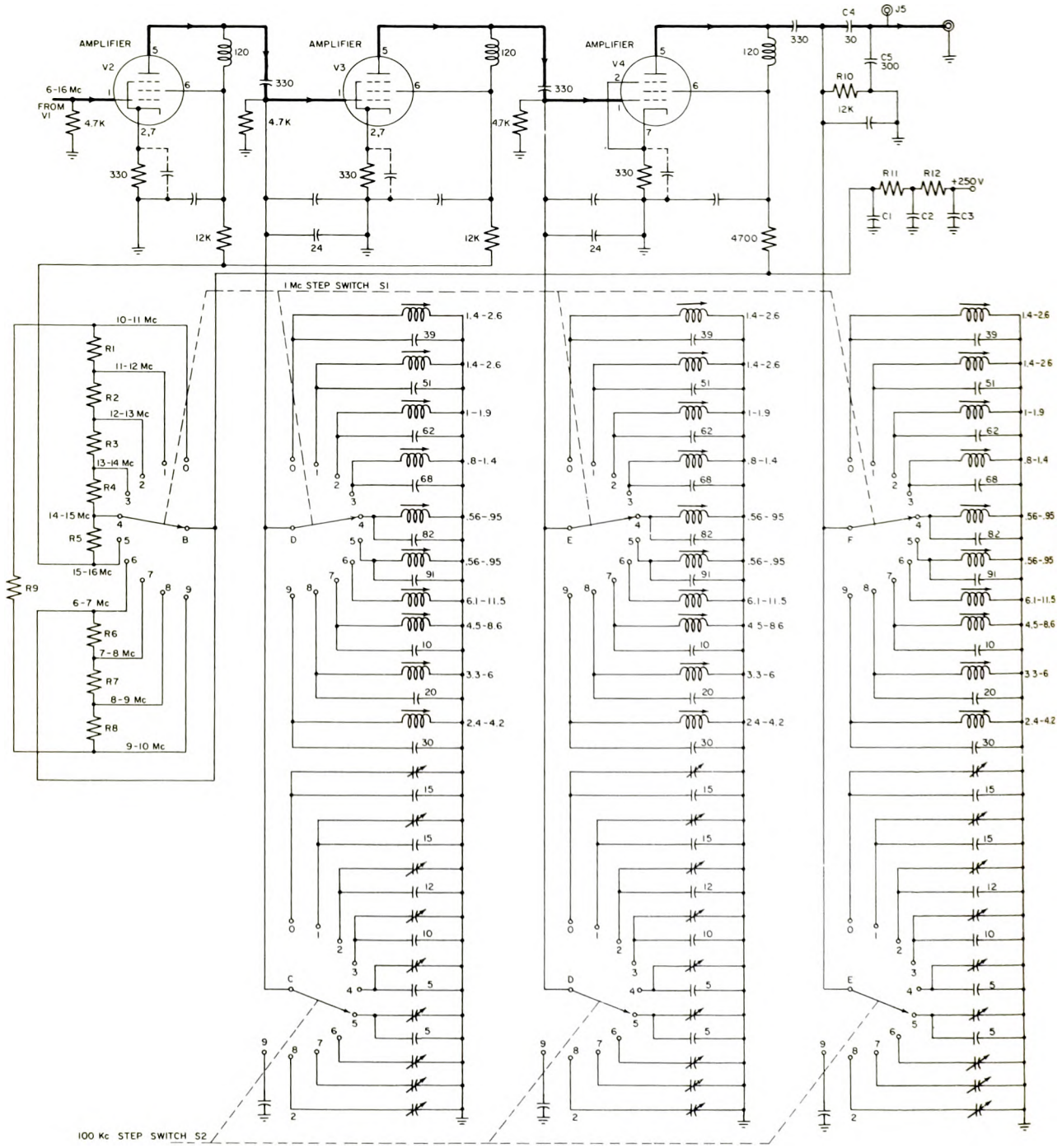

Figure 5-19. –Amplifiers, unit 11B, simplified schematic.

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J1 receives the input from unit 10 and J2 receives the signal from unit 9. Test points J3 and J4 are provided to test the input signal amplitude from units 9 and 10, respectively. The input from unit 10 should be approximately 0.4 volts, and that from unit 9 approximately 1.4 volts.

Unit 11B consists of a mixer, three amplifiers, and switches S1 and S2. The figure representing unit 11B has been divided into two sections. Figure 5-18 shows the mixer and associated components and figure 5-19 shows a similar representation of the amplifiers.

There are nine frequency selection knobs provided for manual selection of the frequency to be transmitted. Two of these controls are used to determine the input frequencies to unit 11. These are: the 5-mc control knob, AA; and the 1-mc control, BB, shown in figure 5-5, B. The remaining seven switches are designated: the megacycle control, GG; the 100-kc control, HH; the 10-kc control, JJ; the 1-kc control, CC; the 100-cycle control, DD; and the 10-cycle control, EE. Control FF (provided for bandswitching) is the bandswitch that has been previously mentioned.

Mixer

The mixer, V1, receives the input from units 9 and 10, which are applied to the suppressor and control grids respectively (fig. 5-18). The frequencies of these inputs are determined by the settings of the controls on the mounting front panel of the radio frequency oscillator, as discussed above.

To select a frequency of 12 mc, paragraph 3 in the discussion of unit 9 shows that knob control AA is to be set at 2. Table 5-1 shows that knob BB should be set at 8. As a result, unit 9 supplies a 32- to 33-mc input and unit 10 supplies a 20-mc input. Mixer V1 combines these inputs to produce an output of 12 to 13 mc.

The 1-mc step switch, S1, selects the proper tuned circuit for the mixer. This switch is controlled by the mc control knob GG. For selection of an output frequency from 6 to 10 mc, control GG is set to the megacycle value of the frequency. For frequencies above 10 mc, control GG is set to the second figure of the megacycle value. The GG control located in unit 11B has no effect during operation from 0.3 to 6 mc because these frequencies are mixed and amplified in unit 11A. Therefore, in selecting a frequency of 12 mc, the GG control is set at 2, and a choice of 7 mc will set the GG control at 7. A table showing the final output frequencies and the AA, BB, FF, and GG control settings is shown at the end of the discussion of unit 11C.

Amplifiers

Amplifiers V2, V3, and V4 are shown in figure 5-19. The amplifiers are identical except that resistors R1 through R9 are provided as dropping resistors for V2 and V3. These resistors are placed in the plate and screen circuits of the tubes through the contacts of the B section of S1. The plate and screen operating potentials are therefore changed at different frequencies, and an approximately equal output is obtained for all frequencies. In the position shown the plate potential is applied through R5. The D, E, and F sections of S1 and the C, D, and E sections of S2 below function in a similar manner to S1 and S2 previously considered in the discussion of the mixer circuit. These switches select the tuned interstage filters for the amplifiers.

R10 is provided to widen the bandpass of the output filter circuit. Unit 11B is decoupled from the power supply by a decoupling filter comprising C1, C2, C3, R11, and R12. Test point J5 is provided for testing the output signal amplitude from unit 11B, which is approximately 2.4 volts.

The final 6- to 16-mc output of unit 11B is coupled by the voltage divider, C4 and C5, to Jack J6, to the bandswitch in unit 14, and to the radio frequency amplifier.

FREQUENCY CONVERTER (UNIT 11C)

This unit is similar to unit 11B. However, it operates in the frequency range from 16 to 26 mc, and therefore requires different component sizes. The keying input from the low-level radio modulator is fed to both the mixer, V1 (fig. 5-20) and the first of the amplifier stages, V2. The step switches, S1 and S2, are mechanically coupled to the step switches in unit 11B.

Mixer

The mixer, V1, receives at its suppressor grid the 7-, 8-, 13-, 14-, 16-, 17-, 20-, or 21-mc input from unit 10. At the control grid the 22 to 23-, 27 to 28-, 32 to 33-, or 37 to 38-mc

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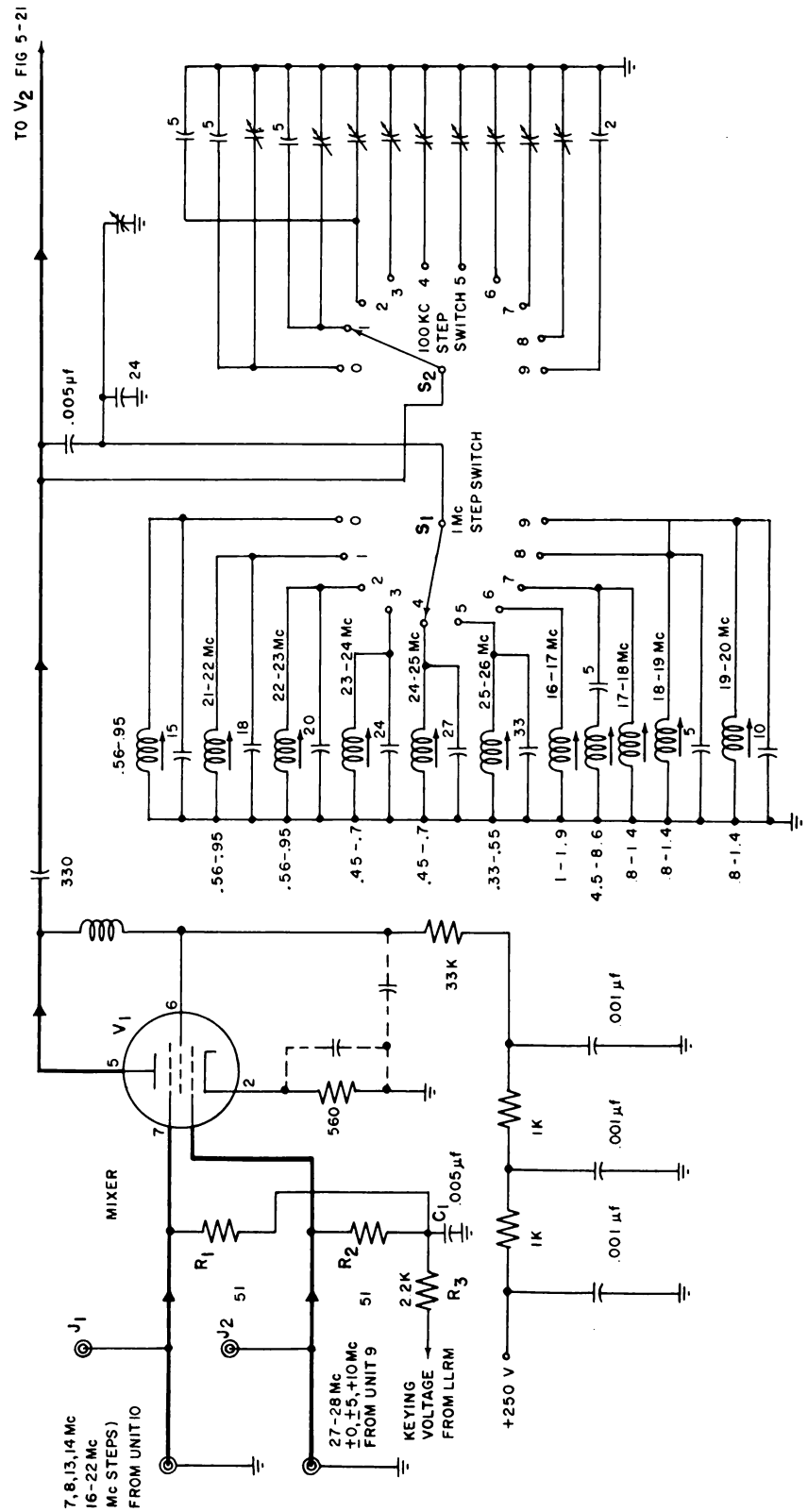

Figure 5-20.—Mixer, unit 11C, simplified schematic.

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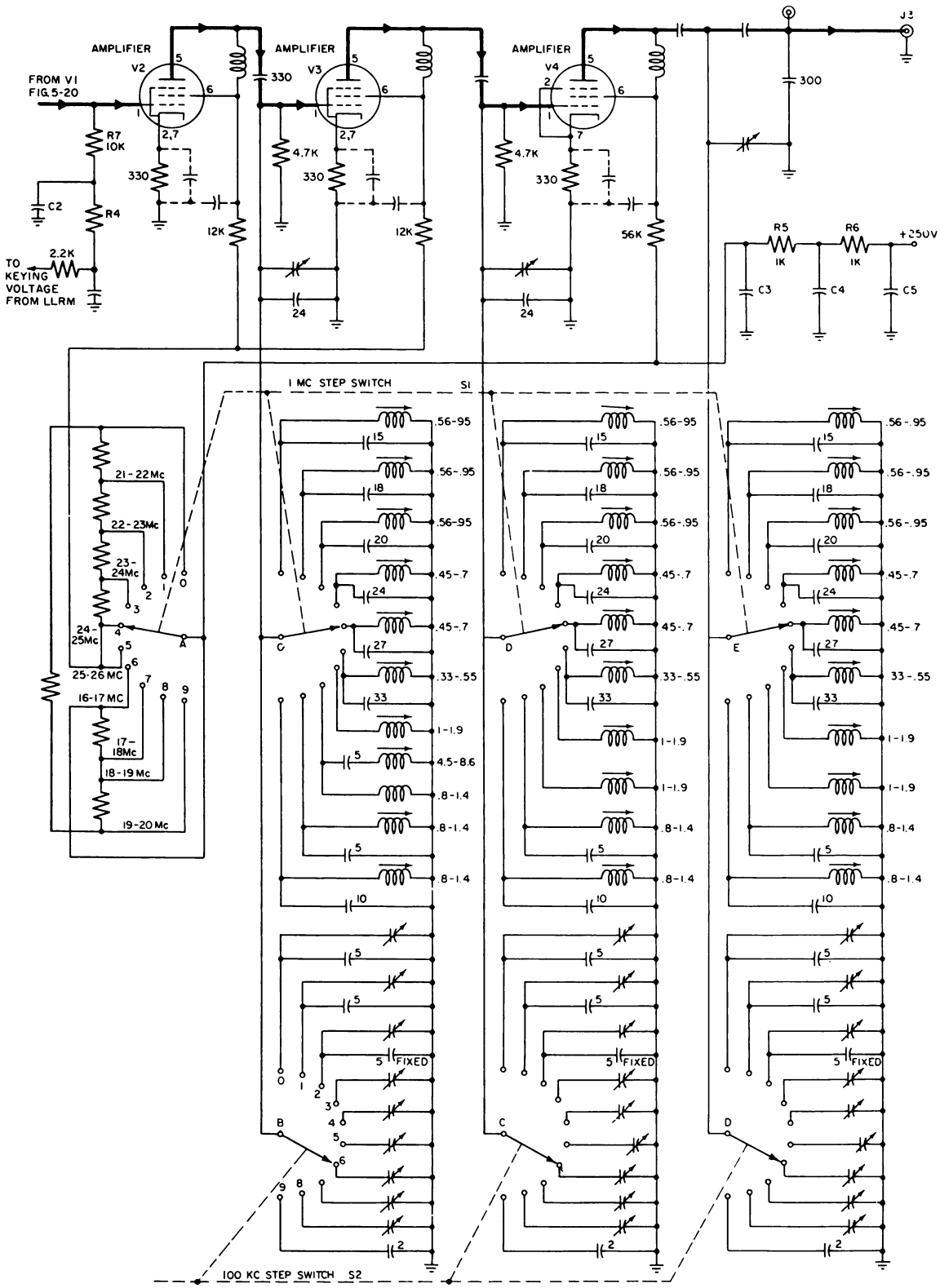

Figure 5-21. –Amplifiers, unit 11C, simplified schematic.

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signal from unit 9 is applied. Test point J1 is provided for testing the input from unit 10, which is approximately 0.4 volts. Test point J2 serves the same purpose for unit 9 input whose amplitude is approximately 1.2 volts. The signals are mixed in accordance with the settings of the AA 5-mc control, S1 (fig. 5-15) and the BB 1-mc step switch, S1 (fig. 5-16). These switches are mechanically connected to S1 and S2 of unit 11C. For example, if the AA and the BB knobs are both set at position 3, unit 9 delivers 37 to 38 mc to unit 11C and unit 10 delivers 14 mc. (Refer to paragraph 3 in the discussion of unit 9 for the frequency settings of the AA control, and to table 5-1 in the discussion of unit 10 for the settings of the BB control.) With these two frequency inputs, mixer V1 delivers a 23 to 24-mc output. Resistors R1 and R2 connect the suppressor and control grids to the low-level radio modulator keying circuit through the filter comprising R3 and C1.

Amplifiers

Amplifiers V2, V3, and V4 (fig. 5-21) are similar to the corresponding amplifiers in unit 11B with the exception that the low-level radio modulator keying is fed to the control grid of V2. The low-level radio modulator keying input is connected first through the same filter network employed in the input to V1, and then through additional filter components R4 and C2 to grid resistor R7. The final 16- to 26 megacycle output of unit 11C, which has been filtered four times is fed to J3 at a level of approximately 2 volts. The B supply for the entire unit is decoupled by C3, C4, C5, R5, and R6.

S1 and S2 control the filter selection for the mixer and amplifier circuits. S1 is the 1-mc step switch, and S2 the 100-kc step switch. As previously stated, these switches are mechanically connected to the switches, S1 (fig. 5-15) and S1 (fig. 5-16).

To illustrate the operation of the controls, assume that an output frequency of 23.57619 mc is desired. To set up this frequency, the 10-cycle control knob EE is set at 9, the 100-cycle control knob DD is set at 1, the KC control knob CC at 6, the 10-kc control knob JJ at 7, and the 100-kc knob HH at 5. These settings have determined the output frequency through the 100-kc level. As discussed earlier, the control AA, BB, FF, and GG are set to determine the megacycle value of the output frequency. Reference to table 5-3 shows that to produce an output frequency of 23 mc, the AA control should be set at 3, the BB control at 3, the FF bandswitch at 16 to 26 mc, and the GG control at 3. Thus, the desired output frequency is produced.

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Table 5-3.--Final Output Frequencies and the AA, BB, FF, and GG Control Positions.

RFO output frequency (mc)	AA control position	BB control position	FF control position	GG control position
0.3-1	0	10	0.3-6 mc	
1-2	0	9		
2-3	0	8		
3-4	0	7		
4-5	0	6		
5-6	1	10		
6-7	0	4	6-16 mc	6
7-8	1	8		7
8-9	0	3		8
9-10	0	2		9
10-11	1	5		0
11-12	1	4		1
12-13	2	8		2
13-14	2	7		3
14-15	2	6		4
15-16	3	10		5
16-17	3	9	16-26 mc	6
17-18	3	8		7
18-19	2	3		8
19-20	1	1		9
20-21	3	5		0
21-22	3	4		1
22-23	0	1		2
23-24	3	3		3
24-25	3	2		4
25-26	2	0		5

QUIZ

1. What is the frequency range covered by the Radio Transmitting Sets AN/SRT-14, 15, and 16?
2. The AN/SRT-14, 15, and 16 provide radio frequency energy for what four types of transmission?
3. What type frequency selection process (manual or automatic) permits all three sets to cover the frequency range in question 1?
4. In the transmitting set AN/SRT-15 (fig. 5-2) the booster provides the necessary facilities for producing what operating output level?
5. What is the master frequency source for the transmitting set?
6. In 500-watt operation, only signals within what range are amplified in the radio frequency oscillator?
7. How many units (fig. 5-5, A) are contained in the radio frequency oscillator?
8. The crystal oscillator unit 1 (fig. 5-5, A) produces a 100-kc output, which is fed to what other units of the radio frequency oscillator?
9. In the frequency multiplier unit 2 (fig. 5-5, A) the 100-kc output of the crystal oscillator is increased to 1 mc and fed to what other units of the RFO?
10. (a) Which unit has an output that is adjustable in steps of 10 cps between 9.19 and 9.2 for c-w and phone, 9.19 to 9.2 ±1000 cps (maximum) for frequency shift telegraphy, and 9.19 to 9.2 2000 cps (maximum) for facsimile operation? (b) To what unit is the output of this unit applied?
11. (a) What is the final output of unit 6 ? (b) to what unit is the output applied?
12. The frequency multiplier, unit 7, receives a 1-mc signal from unit 2 and multiplies it to provide what two frequencies?
13. Which unit serves as a rack into which the other units of the radio frequency oscillator are connected?
14. (a) What stages (fig. 5-6) are included in unit 1 ? (b) which stages in this unit provide a method of coupling the output of the crystal oscillator to the other units of the RFO?
15. The multipliers of unit 2 (fig. 5-7) consists of what two stages?
16. The interpolation oscillator unit 3 (fig. 5-9) generates frequencies in what range?
17. The frequency shift oscillator of unit 12 (fig. 5-10) makes possible what modes of transmission?

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18. If the control tube grid voltage of V2 in the frequency shift oscillator (fig. 5-10) is made less positive, what effect will this have on the oscillator frequency?
19. The output of the frequency shift oscillator unit 12 (100 kc ± $\Delta f/2$) is fed to what unit of the RFO?
20. During c-w and phone operation, what unit is used as the master frequency source instead of the frequency shift oscillator unit 12?
21. (a) How many frequencies are received by the frequency converter unit 5 (fig. 5-11) from other units of the RFO? (b) what are these frequencies and from which units are they applied?
22. What is the function of the blocking oscillator in unit 6?
23. The blocking oscillator frequency located in unit 6 (fig. 5-12) is synchronized to exactly what frequency?
24. To obtain more stable control of the step oscillator of unit 6 (fig. 5-12) the wide pulses from the blocking oscillator are passed through a peaking network comprising what two components?
25. Which component allows only the positive portion of the pulses to appear as the output of the shaper, V4B (fig. 5-12)?
26. The 5-mc signal of unit 7 (fig. 5-13) is fed along dual paths to what two stages contained in this unit?
27. (a) To which unit does the center tap of capacitor voltage divider C3 and C4 (fig. 5-13) couple the 5-mc output of unit 7? (b) the final 15-mc output is coupled to what unit of the radio frequency oscillator?
28. In unit 8 (fig. 5-14) the 1.6- to 2.5 mc output from the phase-locked oscillator is mixed with what other frequency developed in the RFO?
29. What are the frequencies received by unit 9 (fig. 5-15)?
30. The mixer output, V1 (fig. 5-15) is filtered and amplified and sent to any one of which other units of the RFO?
31. What is the position of the switch S1 (fig. 5-15) if the final output of the RFO is 15 mc?
32. What is the name and location of the switch which determines the frequency converter unit (11A, 11B, or 11C) to which the output signals from unit 9 and 10 are delivered and from which units the final output of the RFO will be obtained?
33. (a) When the transmitter is set to operate on a frequency in the range of 0.3 to 6 mc, which of the final frequency converters will be the recipient of the output from units 9 and 10? (b) Which unit is selected when the transmitter is set in the range of 6 to 16 mc? (c) Which unit is chosen when the transmitter is set for a frequency in the range from 16 to 26 mc?
34. Which stage in unit 11A (fig. 5-17) receives the output of unit 10?
35. Why are the two triode section of the cathode follower, V4 (fig. 5-17) connected in parallel?
36. (a) What frequencies does unit 11B (fig. 5-18) receive from unit 10? (b) what frequencies are received by this unit from unit 9?
37. How many frequency selection knobs are provided for manual selection of the frequency to be transmitted?
38. (a) For selection of an output frequency from 6 to 10 mc, how is the GG control set? (b) for frequency selection above 10 mc, how is this control set?
39. The keying input from the low-level radio modulator is fed to which stages of unit 11C (fig. 5-20 and 5-21)?

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CHAPTER 6

OPERATING PRINCIPLES OF A REPRESENTATIVE RADIO TRANSMITTER, PART II

INTRODUCTION

Operating Principles of a Representative Radio Transmitter, Part II deals primarily with the circuitry of the AN/SRT-14, 15, and 16 transmitters. These units were chosen for presentation in this chapter because their circuits represent, to a great extent, all of the new features of present day transmitters. The AN/SRT-14, 15, and 16 are similar in circuitry; they differ mainly in their relative power output levels. Several booster circuits are contained in the higher power transmitters to provide the required output power level.

The design of these transmitting sets is unique in many ways. It is assumed that you have read and understood the material presented in Chapter 5, Operating Principles of a Representative Radio Transmitter, Part I, and that you can now understand the new application of heterodyning in the units of the RFO to produce the desired output frequency. It was explained in that chapter that through the heterodyning process, the main oscillator frequency was permitted to remain unchanged. This improves the stability of the output frequency and, likewise, the quality of the transmitter.

In this chapter, particular emphasis is placed on the radio frequency amplifier and associated antenna equipments of the AN/SRT-14, 15, and 16. While studying this chapter, you should relate the major units of this transmitter to earlier ones with which you are more familiar.

GENERAL DESCRIPTION OF THE TRANSMITTER

All transmitters employ some combination of oscillators, frequency multipliers, power amplifiers, and audio circuits, which produce and modulate the generated radio frequency carrier with an intelligible signal. Many more advanced transmitters make use of several additional circuits, including mixers, step generators, standing wave ratio monitors, and special keying circuits, which provide great improvement in the quality of the output and the distance a given message may be transmitted.

The transmitters probably most familiar to you include the TBL, TBK, and TDE, which are, at this time, obsolete or obsolescent. Although these transmitters are being removed from most ships, their replacements are in many respects the same. In addition, they incorporate new circuits and concepts that improve the overall efficiency.

The prime function of the radio frequency amplifier of the AN/SRT-14, 15, and 16 transmitters is to increase the power of the r-f output to the required level. Each unit is mounted on a separate chassis. After studying the RFO (chapter 5), it is easy to see that the space required for such a unit is quite large. By placing the units on separate chassis, the amount of interaction between frequencies of the two units is minimized. A third factor that contributes to the choice of separate chassis for the RFO and RFA (and other major units of the transmitter) can be appreciated when it becomes necessary to service the equipment. For instance, if an operational trouble is found in the RFA, this unit alone may be replaced by removing the RFA chassis and substituting another, which is known to be in operating condition.

In addition to amplifying the r-f energy, the RFA is equipped with tuned circuits and filters, which are selected in the plate circuits through the use of switches provided for this purpose. A decided improvement is therefore realized over earlier transmitters in the ability of this equipment to select the desired frequency. Also, since each of the six bands utilizes a separate filter, the range of tuning of a particular tank circuit is minimized.

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The output circuits of the RFA contain meters, switches, and reactive components, the latter being used to match the RFA to the chosen antenna. Through the use of the transmitter coupler, r-f tuner, and antenna coupler units of the RFA, the electrical length of the transmission line can be varied for matching to a fixed antenna, and the standing wave existing along the line can therefore be reduced to the required minimum. These features greatly improve the distance and quality of the transmission. A block diagram of the AN/SRT-14, 15, and 16 transmitter is shown in chapter 5 (fig. 5-4).

RADIO FREQUENCY AMPLIFIER

The radio frequency amplifier (RFA) accepts from the radio frequency oscillator a signal in the frequency range from 0.3 to 26 mc. The radio frequency amplifier consists of three stages: the buffer, the intermediate power amplifier (IPA), and the power amplifier (PA). The signals from the radio frequency oscillator are received at a level of approximately 0.1 watt, and amplified to either 100 or 500 watts. The radio frequency amplifier also contains adequate filters to attenuate undesired harmonics. The output of the RFA is coupled to the load-adjusting unit.

The RFA output may be either modulated by audio from the low-level radio modulator or high-level radio modulator, or it may be keyed on and off by keying voltages from the low-level radio modulator. During frequency-shift keying and facsimile operation, the r-f signal from the RFO is frequency modulated in the radio frequency oscillator. The keying circuit are considered in the discussion of the low-level radio modulator, which appears in chapter 7 of this text.

The Tuning of the RFA is accomplished manually by three controls: the bandswitch, the tune IPA control, and the tune PA control. To enable the RFA to tune through the entire range from 0.3 to 26 mc, the total range of frequencies has been grouped into six bands. Figure 6-1 shows a block diagram of the RFA with its ganged bandswitches. For the sake of simplicity, meters and control circuits are not shown.

The function of the bandswitch is to connect a tank circuit to each of these amplifiers in accordance with the band that contains the frequency chosen for transmission. After the desired band is chosen by the bandswitch, the tuning of the buffer and IPA tank circuits is accomplished by the tune IPA control. The

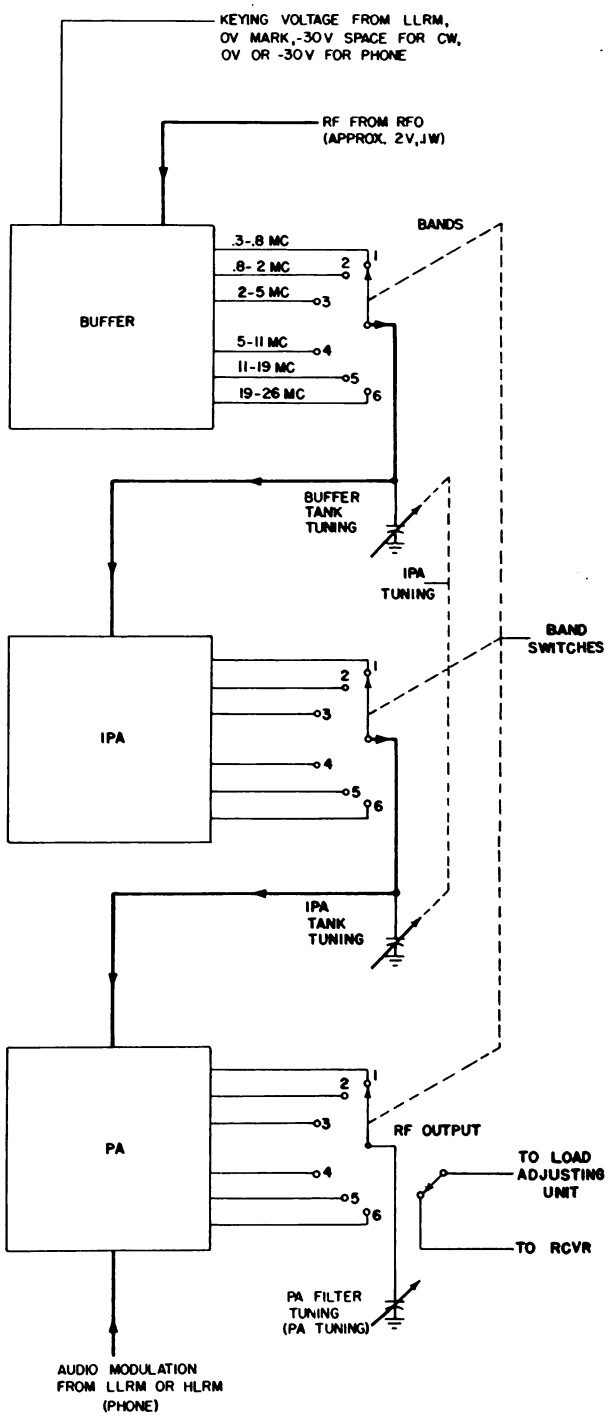

Figure 6-1. –Radio frequency amplifier, block diagram.

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tuning of the PA tank circuit is then performed by the tune PA control. Common operating adjustments of the AN/SRT-15 transmitter are described in Chapter 12 of *Electronics Technician 3,* NavPers 10188-A.

The function of the buffer stage is to isolate the RFO from the effects of a varying load caused by keying signals received by the RFA from the low-level radio modulator.

BUFFER AMPLIFIER

The buffer amplifier circuit (fig. 6-2) comprises tube V1. This circuit, operated class A, receives the output of the RFO and amplifies the signal to a sufficient value to drive the intermediate power amplifier. The output of the radio frequency oscillator is terminated by excitation control R1. This control, a 100-ohm potentiometer, allows impedance matching of the coaxial cable carrying the r-f signal from the radio frequency oscillator to the input circuits of the RFA. When the impedance input is correctly adjusted, maximum power transfer will exist between the RFO and the RFA, and the same signal strength will be presented to the RFA input regardless of the length of the line connecting the two circuits. The final setting of the excitation control should allow an input to the buffer, V1, not exceeding 5 volts rms, or should be such that when all tuning is completed, the value of the PA screen grid, V3, current does not exceed 70 milliamperes. The input to the buffer, V1 passes through C1 and the INT position of S1.

The selection of the V1 plate tank circuit components is determined by the setting of the ganged bandswitch sections, S2A, S2B, and S2C. The switches are shown in band 1 position. To locate the components comprising the plate circuit of the buffer with the bandswitch in this position, trace from the plate of V1, to the continuous contact of S2C rear, to contact 1 of this switch, through L1 (paralleled by R2) through L2 (bypassed by C2), to the B supply. Tuning capacitor C3 is connected across the inductor, L1, and damping resistor R2. R2 broadens the band response of the tank circuit which is necessary to maintain even excitation during frequency-shift operation. Resistors R3, R4, R5, and R6 perform a similar function for their respective bands. Inductors L3, L4, L5, L6, and L7 are the plate tank inductors for the remaining bands.

S2B shorts out all unused inductors. C3 connected by S2A parallels the chosen plate inductor and in bands 5 and 6 selects band spread capacitors C4 or C5 respectively. The addition of either C4 or C5 in series with C3 reduces the tank tuning capacitance and spreads the tuning of bands 5 and 6 over the full 180-degree rotation of the tuning capacitor. After the desired tank circuit of the buffer has been selected, it may be tuned to maximum output by varying capacitor C3, which is manually controlled by the tune IPA control. The output of the buffer stage is coupled by C6 to the control grid of the IPA, V2. C7 is a trimmer capacitor for the V2 input circuit.

Zero volts (ground) is applied to the buffer control grid during frequency-shift keying and facsimile operation, through R7, from the low-level radio modulator (discussed later). The screen potential is obtained from the regulated +150 volts supplied by the action of the voltage regulator tube, V4. During frequency-shift keying and facsimile operation, the buffer amplifier functions to key the carrier continuously on the air. In c-w operation during mark and in phone operation when the press-to-talk button is depressed, the signal from the LLRM (chapter 7) is such that the buffer will also allow the carrier to be transmitted. However, in c-w operation during space and in phone operation when the press-to-talk button is not depressed, a minus 30 volts is received from the low-level radio modulator. This potential is applied through R7 to the control grid of the buffer, and is sufficient in amplitude to drive the stage to cutoff. In this condition, the buffer amplifier will not allow passage of the r-f signal from the RFO and effectively cuts the carrier off the air. The origin of the 0- and -30-volt keying signals will be further considered in the discussion of the low-level radio modulator.

INTERMEDIATE POWER AMPLIFIER

The potential on the voltage divider (in the V2 grid circuit) between the R8-R9 junction and ground is -65 volts. The circuit connected to the -220-volt power supply is grounded through the antenna tuning unit bypass at the junction of R9 and R10. The -65 volts is applied to the V2 intermediate power amplifier control grid. In addition, a grid leak circuit composes of r-f choke L8, R11, and C6 also provides a bias voltage for this stage. Thus, a combination of fixed and grid leak bias is used.

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If the grounding line through the antenna tuning unit is interrupted, as will be the case when this unit is switched out, the resistor, R10 (in the control grid circuit) will be added to the voltage divider. The addition of this resistor will increase the bias of the intermediate power amplifier from -65 volts to approximately -180 volts. Thus any tendency for an r-f surge in the following stage caused by the momentary loss of antenna load will be counteracted by the increased IPA bias. The grounding line to the antenna tuning unit is further considered in the discussion of the antenna tuning equipment, which appears later in this chapter.

R12 is a grid current meter shunt bypassed by capacitor C8. The cathode is grounded through a parasitic suppressor, R13, and a cathode meter shunting resistor, R14, the latter resistor being bypassed by capacitor C9.

The screen grid circuit contains parasitic suppressor R15, screen filter resistor R16, and a screen current meter shunting resistor R17. R18 and R19 are meter multiplier resistors. The plate circuit contains E1, which is also a parasitic suppressor. This arrangement of suppressors is common to many transmitter circuits and is found particularly in stages whose input and output circuits are tuned to the same frequency.

The buffer amplifier, V1, and the intermediate power amplifier, V2, are tuned by the tune IPA control, which is ganged to the IPA tuning capacitor, C10, as well as to the buffer tuning control, C3. The desired tuning inductor and associated damping resistor are selected by S2D and S2F. All other tuning inductors and damping resistors are shorted by S2E. On bands 5 and 6, bandswitch S2D selects either bandspread capacitor C11 or C12, respectively, which is placed in series with the tuning capacitor, C10. The r-f output of the IPA, approximately 210 volts rms, is coupled by C13 to the control grid of the power amplifier, V3. The V2 IPA plate circuit is decoupled from the +500-volt power supply by L9 and C14.

The power amplifier receives the r-f output from the intermediate power amplifier. When a key input is applied to the transmitter, the key-up radiation must be maintained below a maximum allowable level of 400 $\mu\mu$w. Therefore, in addition to a higher plate and screen voltage required for 500-watt operation as compared to 100-watt operation, a higher bias voltage is also needed to maintain the desired level of the key-up output. The two individual bias voltages are obtained from two taps on a voltage divider network, comprising R20 through R23, connected from the -220-volt supply. In 100-watt operation, grid bias is obtained from the tap on potentiometer R20. The bias voltage is applied to the control grid of the PA, through contacts 5R and 6R of relay K1, through the grid current meter shunting resistor R28, grid return resistor, R29, and r-f choke L23. Relay K1 is not energized during 100-watt operation. In 500-watt operation, the relay is energized and grid bias is taken from a more negative point on the divider (the tap on potentiometer, R21). This potential is then applied to the PA control grid through contacts 10L and 9L of the relay. The potentiometers, R20 and R21, are normally adjusted for -110 and -155 volts, respectively.

POWER AMPLIFIER

The output circuits of the power amplifier, V3 (fig. 6-2) are shown in the band 1 position. The output circuit is essentially a shunt-fed L-C tank, which acts as a harmonic filter, the the components of which are determined by the setting of the bandswitch. All of these switches are set by a single bandswitch control. Since the buffer and IPA are controlled by sections of the bandswitch, the circuits of the three stages are changed simultaneously. S2G selects the r-f chokes in the PA plate supply circuit across which the r-f voltage is developed. To keep the r-f voltage level constant for all frequencies, more inductance is required at the lower frequencies. The chokes operate at the resonant frequency created by their inductances and their distributed capacitances in order to present a high shunt impedance to the V3 plate and the output tank. Therefore, on band 1, all of the chokes, L10, L11, L12, L13, and L14, are connected in series in the PA plate circuit. On band 2, L10 is shorted by S2G. On band 3, L11 and L10 are shorted, and so on until on band 5, only L14 is placed in the plate circuit. Inductor L15 and capacitor C14 form a decoupling network for the plate circuit of the PA. The two components are resonant just above the center of the frequency range covered by the transmitter. In this manner, the decoupling process is improved. Capacitors C15 and C16 suppress any tendencies of the chokes to resonate under the influence of the audio modulating signal impressed on the plate circuit during phone operation. Switch sections S2H and S2J

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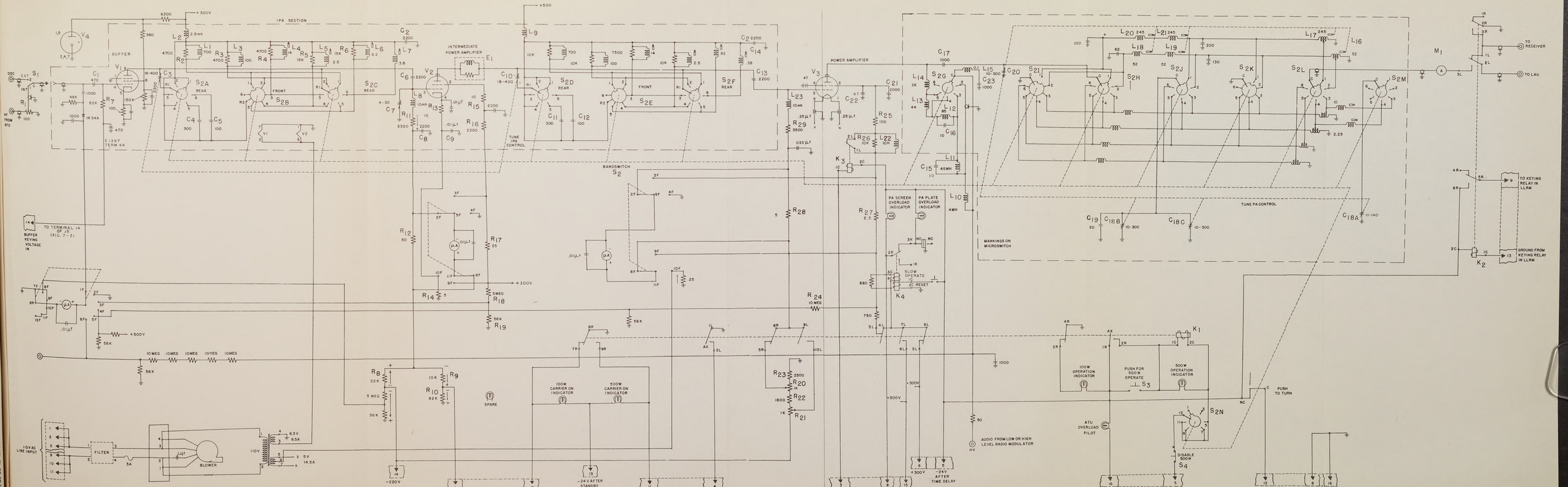

Figure 6-2. –Radio frequency amplifier, schematic diagram.

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select the required tuning inductors, while S2I and S2K short out the unused inductors. S2J also selects the required antenna matching inductor and connects capacitors C18A, C18B, C18C, and C19 in parallel with it. Switch S2L couples the output from the impedance matching coils through the antenna current ammeter, M1, to contact 3L of the antenna switching relay, K2. If K2 is energized, the output of the PA is fed, along a 50-ohm line, to the load-adjusting unit, through the relay contacts, 3L and 2L. When K2 is not energized, the antenna system is connected to the 1L and 2L contacts, which may be applied to a receiver. Proper switching of the antenna will allow its use with a receiver when and if a receiver is used. A complete analysis of the antenna switching to the receiver is given at the end of the discussion of the RFA. Switch section S2M provides additional impedances as required to the antenna matching inductors in bands 3 and 4.

After the proper components of the filter have been selected, the filter is tuned to maximum output. The tuning control simultaneously adjusts C18A, C18B, and C18C and inductors L16, L17, L18, L19, L20, and L21.

During c-w transmission, the V3 PA plate voltage is +1300 volts for 100-watt operation and +3000 volts for 500-watt operation. During phone transmission, a 1050-volt potential is supplied to the V3 plate for 100-watt operation and 2400 volts for 500-watt operation. The plate potential is reduced during phone operation so that during modulation the maximum plate variations will not exceed the V3 peak voltage ratings. The audio modulation voltage is inserted in series with the V3 plate supply.

In 100-watt operation of the transmitter the 300-volt source supplies the PA screen, while in the 500-watt operation the 500-volt potential is applied to the screen. Relay K1 provides the necessary switching, being energized for 500-watt operation and deenergized in 100-watt operation. The 300-volt supply is connected in 100-watt operation through contact 5L and 6L of K1. The r-f signal in the screen circuit is filtered by R25, C21, and C22. The latter capacitor also tends to suppress parasitic oscillations. When the modulation voltage is introduced in the plate circuit of V3, it is also necessary to introduce a portion of the modulation voltage in the screen circuit to achieve 100 percent modulation. Audio signals in the screen circuit are produced across the audio choke, L22, as a result of the audio plate variations injected in series with the plate supply during phone transmission.

For example, on the positive half of the audio cycle when the plate current is periodically increasing, the screen current is decreasing. (For a given filament temperature the space current is constant.) This action decreases the drop across the audio choke, L22, in the screen circuit and automatically raises the screen voltage the necessary amount to accomplish 100-percent modulation of the r-f carrier in the shunt-fed plate circuit. Conversely, on the negative half of the audio cycle when plate current is decreasing, screen current is increasing, and the drop across L22 increases the amount of voltage subtracted from the screen supply to lower the screen voltage accordingly. This action further decreases the plate current pulses to accomplish 100 percent modulation of the negative half of the audio cycle.

Therefore, the r-f energy in the screen circuit is bypassed, while the audio variations are impressed at the screen grid by the modulation network comprising L22 and R26. The corresponding boost in the audio signal amplitude causes the percent of modulation in the PA to be increased. This action is necessary for phone operation only, and relay K3 functions such that the modulation network is shorted in all modes of operation except phone. During phone operation, K3 is energized from the -24 VOLTS AFTER TIME DELAY to the GROUND FOR PHONE line, opening the contacts, 1L and 2L, which removes the short around L22 and R26. R27 acts as a meter shunt. The +500-volt supply is applied to the V3 screen grid in 500-watt operation through contacts 7L and 8L of the energized K1.

500 W OPERATE RELAY AND 500W DISABLE LINE

The 500-watt operate relay, K1 (fig. 6-3) controls the GROUND FOR 100 W line, the GROUND FOR 500 W line, and the GROUND FOR 500 W AX line.

The GROUND FOR 500 W line and GROUND FOR 500 W AX line allow circuit completion for several control circuits. They are responsible for turning on control circuits in the high voltage power supply, for connecting the output of the low-level radio modulator to the input of the high level radio modulator, for switching the input to the PA plate from the low-level radio modulator output to the high-

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level radio modulator output, and for supplying screen grid voltage to the high-level radio modulator during phone operation. The GROUND FOR 500 W line and GROUND FOR 500 W AX line are completed when K1 is energized. The completion is made through contacts 8R and 9R, and 1L and 2L, respectively, of K1.

The GROUND FOR 100 W line is completed when K1 is not energized. This line completes the control circuits in the medium voltage power supply, which then supplies the required PA plate potential for 100-watt operation.

If K1 is energized, the transmitter group will be in 500-watt operation. If the relay is not energized, the transmitter will be in 100-watt operation. To place the transmitter in 500-watt operation, the Push for 500-Watt button, S3, is depressed. If the -24 volts after time delay relay, K5 (to be discussed later) has timed out (contacts 2-8 closed), depressing the 500-watt button will apply the -24 volts to the winding of K1. The necessary ground for completion of this circuit is accomplished through a series of protective devices, constituting the 500 W disable line. The circuit must

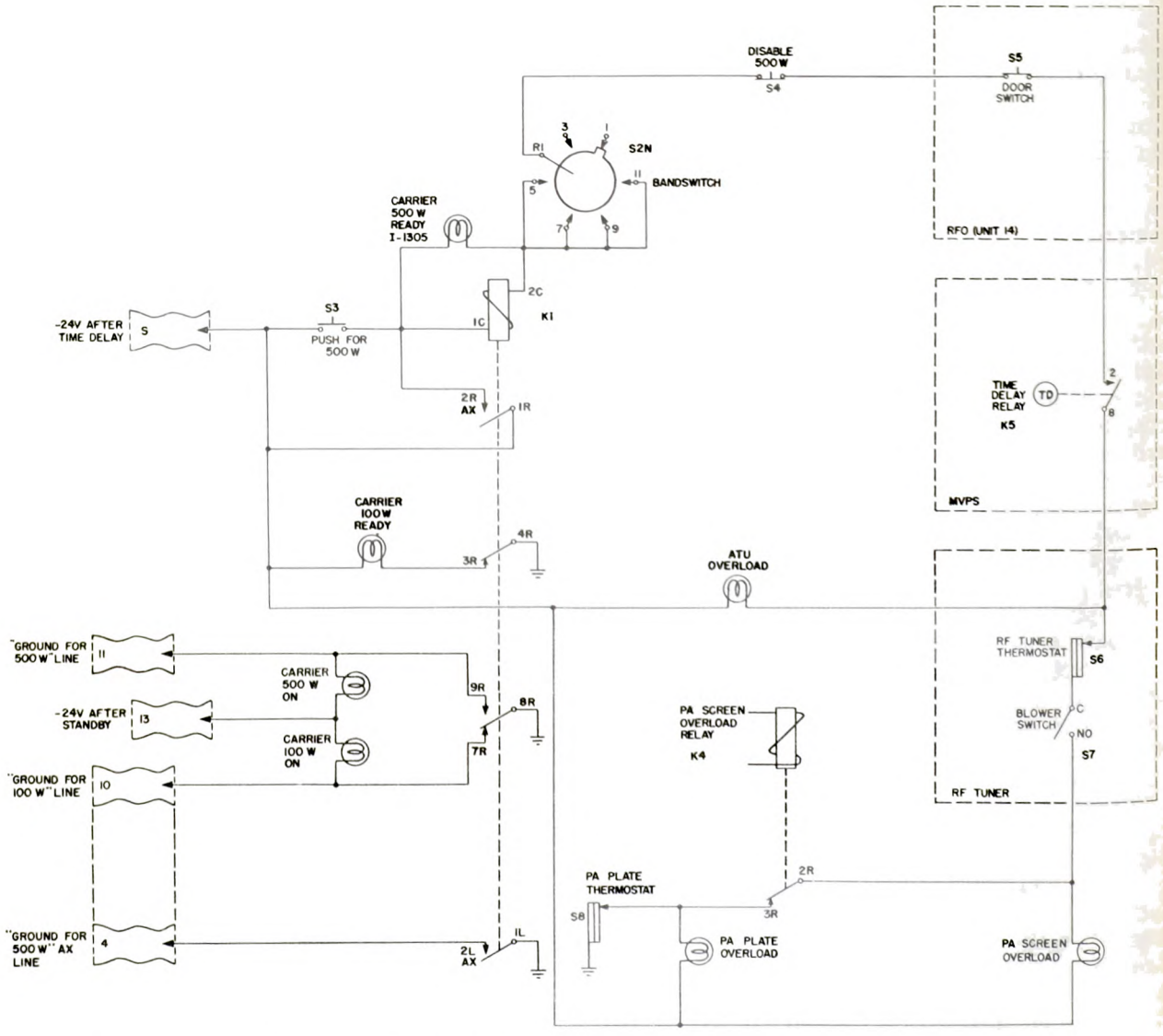

Figure 6-3. –500 Watt Operate Relay and 500 Watt Disable line, simplified schematic.

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be complete before the transmitter will switch to the 500-watt level. The protective devices included in the 500 W disable line are the bandswitch, S2N; the disable 500 W pushbutton, S4; the RFO unit 14 door interlock, S5; the time delay relay, K5; the thermostat, S6; the blower switch, S7; the overload relay, K4 (contacts 2R and 3R); and the plate thermostat, S8. If any one of these controls or protective devices is not closed, the ground will not be completed to K1, and the transmitter can operate at the 100-watt level only. When K1 energizes, -24 V after time delay connects to the windings of K1 through its contacts, 1R and 2R, allowing S3 to be released.

The 500 W disable line contains a switch, S2N, which is ganged to the bandswitch control. The switch allows the circuit to be completed in all bands except 1 and 2. It should be recalled that those frequencies up to and including 2mc may be transmitted at the 100-watt level only.

The disable 500 W pushbutton, S4, which is normally closed, may be depressed to open the circuit as desired.

A transparent door covers the control knobs of the RFO. The door contains a switch, S5. This door interlock controls the open and closed condition of the 500 W disable line, and, if open during normal 500-watt operation, the transmitter output will be reduced to the 100-watt level.

The heaters of the tubes in the high-voltage power supply (discussed later) are allowed to warm up before power is applied to the plate transformer. The contacts 2 and 8, of the time delay relay, K5, are held open until the time delay is over. After this time, the contacts 2 and 8 are closed to complete the circuit.

Thermostat S6 is normally closed. When excessive heat is radiated by tuning coils in the r-f tuner (discussed later in this chapter), S6, will open the circuit.

Blower switch S7 is a centrifugal switch operated by the blower motor in the r-f tuner. The circuit is closed through S7 only if the blower is operating.

The contacts, 2R and 3R, of the PA screen overload relay are normally closed. When K4 is energized, due to excessive PA screen grid current, contacts 2R and 3R are opened, and the ground is removed from the 500 W disable line.

The normally closed contacts of the PA plate thermostat, S8, are caused to open when the energy radiated by PA plate dissipation becomes excessive. This, too, will remove the ground from the 500 W disable line.

Four indicator lamps controlled by the action of K1 indicate the power level of the transmitter and also whether or not the transmitter is on the air. These indicators are shown in figure 6-3.

As mentioned previously, contacts are provided on the antenna switching relay, K2 (fig. 6-2) for connecting a receiver to the antenna when the transmitter is off the air. When the transmitter is on the air, the output of the RFA is coupled to the antenna system, and the receiver input is grounded.

During hand-key operation (using the test key, S5 in figure 6-4) the circuit for K2 is completed from the -24 v after time delay, through the winding of K2, into the low-level radio modulator, through contacts 1L and 2L of relay K6, and through the closed contacts of the test key to ground. Energizing the antenna switching relay, K2, actually operates the keying relay, K6, in the LLRM. This action shifts the ground on K2 from 1L and 2L of K6 to the contacts 2L and 3L of K6. C1 is provided to continue the ground on K2 during the time of the changeover. The antenna relay, K2, therefore becomes energized when the test key is closed corresponding to a mark, moving its contacts, 3L and 2L (fig. 6-2) to the transmit position. When the hand-key is up, corresponding to a space, there is no ground for relay K2 and the antenna is switched to the receive position.

In machine c-w, frequency-shift keying, and facsimile operation, a permanent ground is placed on the keying line (through the keying equipment), which keeps K2 energized.

TRANSMITTER COUPLER (LOAD ADJUSTING UNIT, LAU)

The function of the load adjusting unit (fig. 6-5) is to improve the impedance match between the 50-ohm output impedance of the radio frequency amplifier and the impedance presented by the antenna tuning equipment. The improved matching permits the final PA stage of the radio frequency amplifier to operate at maximum plate efficiency, and improves the transfer of power to the antenna system. The load adjusting unit consists of the standing-wave ratio monitor and the impedance transformer circuits.

Standing Wave Ratio Monitor

The output of the radio frequency amplifier is connected to the standing-wave ratio monitor

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circuit, which functions to detect and monitor the standing-wave ratio present on the r-f transmission line. The components together of the standing-wave ratio monitor circuit form a unit designated the directional coupler. A diagram of this unit is shown in figure 6-5.

The r-f transmission line from the RFA is connected through J1 to the standing-wave ratio monitor circuit and to the impedance transformer circuit. Voltages eV_1 and eV_2, proportional to the r-f line voltage, are obtained directly from the voltage divider, R1, and d-c blocking capacitors C1 and C2, respectively. Voltage eV_1 is between point C and ground. Voltage eV_2 is between point B and ground. Voltages ei_1 and ei_2 proportional to the current in the r-f line, are obtained from the lower and upper halves, respectively, of the secondary winding of the transformer, T1. The center tap of the secondary is grounded, and ei_1 is 180° out of phase with ei_2. However, eV_1 and eV_2 are in phase with each other and are of equal magnitudes. Resistors R2, R3, R4, and R5 shunted across the transformer function as damping resistors.

Circuit ABCD supplied by the T1 secondary and also by a portion of the line voltage via C1 and C2 includes capacitors C3 and C4, which acquire a charge through the action of crystal diodes CR1 and CR2 and the voltages ei_1, eV_1 and ei_2, eV_2.

On one-half cycle, CR1 and CR2 are conducting, and on the other half cycle they are nonconducting. During the time the diodes conduct, ei_1 and ei_2 attempt to establish a charging current clockwise around circuit ABCD so as to charge C3 and C4 with the polarity shown. During this time eV_1 opposes the charge on C3, and eV_2 aids the charge on C4.

As stated before, ei_1 and ei_2 are proportional to the line current; eV_1 and eV_2 are proportional to the line voltage. Now the quantity of charge in C3 is proportional to $ei_1 - eV_1$; the quantity of charge in C4 is proportional to $ei_2 + eV_2$. Thus, the voltage on C3 is proportional to the difference of two quantities, and C4 is proportional to the sum of the same two quantities. These quantities are proportional to line current and line voltage.

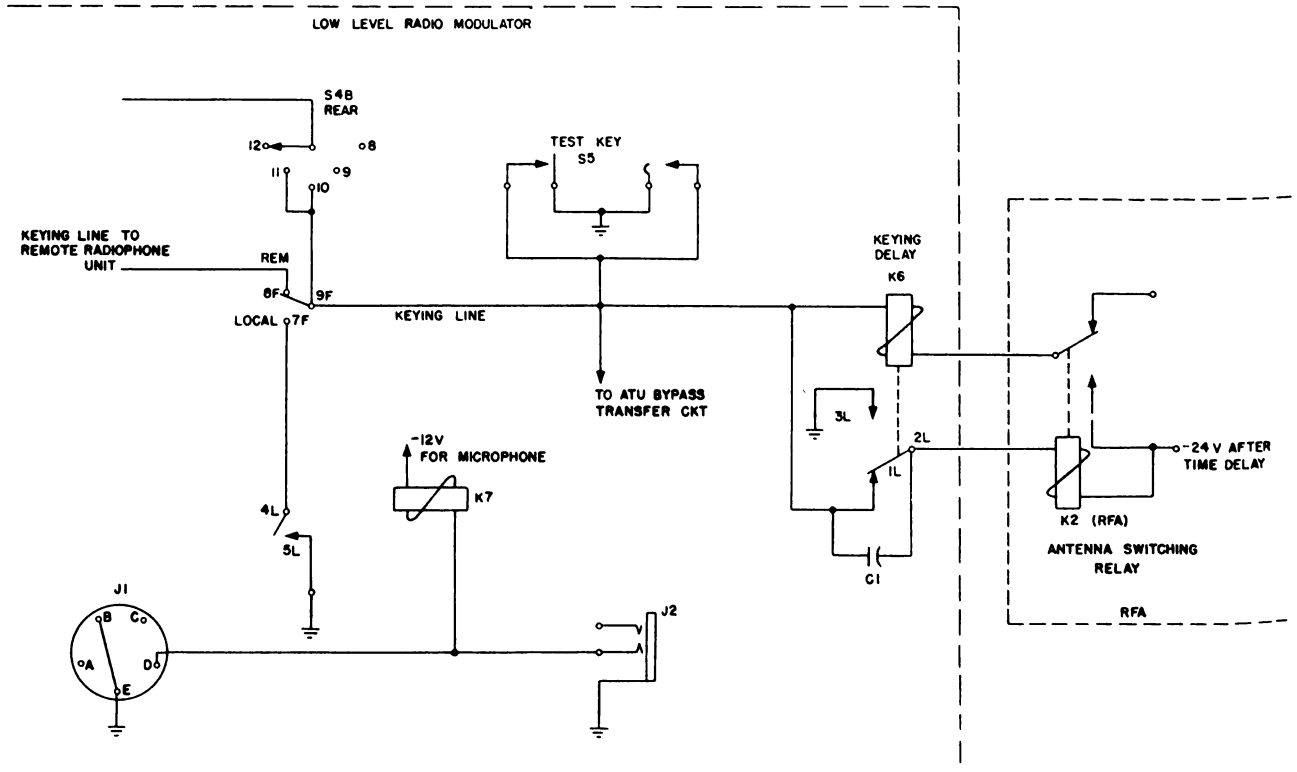

Figure 6-4. –Antenna switching relay and associated circuits.

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The ratio of these quantities is proportional to the line impedance; for a matched line with no standing waves, the ratio is proportional to the characteristic impedance of the line. If the line is mismatched, the ratio is upset, and the amount of charge on C3 and C4 is altered.

A voltage divider comprising R8, R9, R10, R11, R12, and R13 is connected across the series combination of C3 and C4. The voltage across this divider is equal to the sum of the voltage across C3 and C4. The meter, M, is connected effectively between the common junction of C3 and C4 and a point on the voltage divider, depending on the position of switch S3. In this bridge circuit the current through the meter is proportional to the unbalanced condition of the bridge.

The values of R10, R11, R12, and R13 are such that if the standing wave on the r-f line is 8:1, the potential at point E with respect to ground is zero. If the SWR CALIBRATE switch S3, is in the 8:1 position, the SWR BALANCE meter, M, would give an indication at the center null position. If the standing-wave ratio is less than 8:1, with the switch in this position the meter will read to the left of the null position. Conversely, if the standing-wave ratio is greater than 8:1, with the switch in

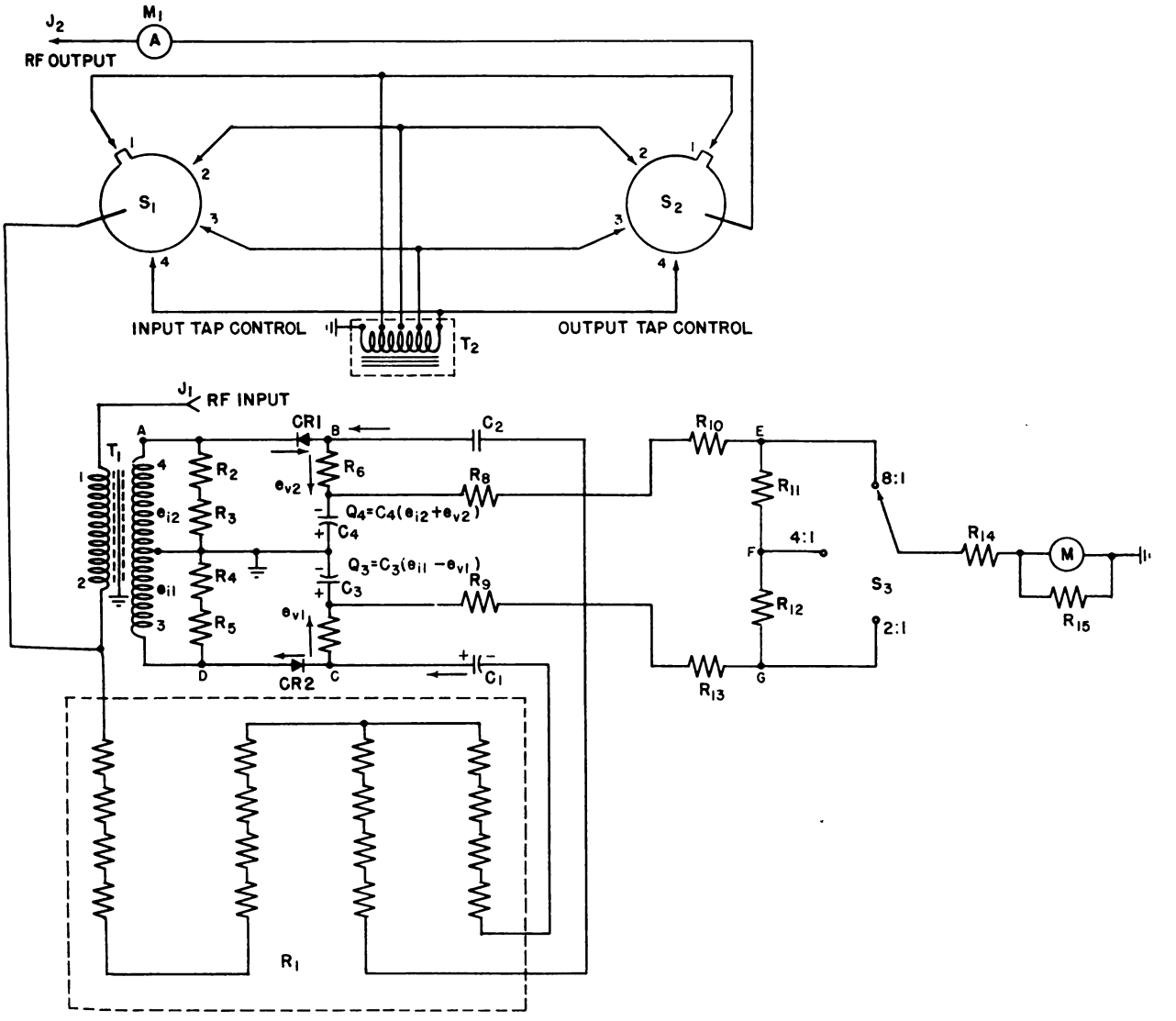

Figure 6-5. —Transmitter coupler (load adjusting unit), schematic diagram.

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this position the meter would indicate a reading to the right of the null position. The voltage at point F with respect to ground will be zero when the standing-wave ratio is 4:1. Similarly, a 2:1 ratio will give a zero reading at point G. R14 in series with the meter is a current-limiting resistor.

Impedance Transformer

A method for improving the match between the characteristic impedance output of the RFA and that impedance presented to the RFA by the antenna and antenna tuning equipment is provided through the use of the impedance transformer. Matching is accomplished through the use of a tapped autotransformer T2 (fig. 6-5), which has been introduced into the r-f transmission line. The output of the standing-wave ratio monitor may be connected to one of the taps of T2 shown in figure 6-5. The input connections are made through switch S1, the input tap control. The output of the transformer selected by the position of the tap on the output tap switch, S2 is connected through the r-f ammeter M1, to the jack, J2, which is the output to the antenna coupler.

With the input and output tap controls set at position 4, the full winding of the impedance matching autotransformer will be shunted across the line. This is the normal setting of these controls. If, after all tuning procedures have been completed, the standing-wave ratio balance meter indicates a reading higher than 2:1, the controls should be reset for optimum impedance match. The tuning controls of the PA stage are also used in making the adjustment for maximum reading on the r-f ammeter of the load adjusting unit.

ANTENNA TUNING EQUIPMENT

The AN/SRT-14, 15, and 16 transmitting sets are designed to work into a 35-foot whip antenna Navy type C66047 or into a 60- to 130-foot single wire with a 40-foot down lead. If the impedance presented by the antenna is other than 50 ohms, energy will be reflected back along the transmission line, reducing the energy being radiated and causing voltage nodes to exist on the line. The standing-wave ratio measurement indicated by the standing-wave ratio balance meter represents the extent of the mismatch. The ideal matching condition of the antenna load to the RFA output would yield a standing-wave ratio of 1:1. This would occur when the reactive component of the antenna is reduced to zero and the resistive component is equal to 50 ohms. Acceptable tuning of the antenna load by the antenna tuning equipment is achieved if the standing-wave ratio can be reduced to at least 4:1 for any frequency in the transmitter range.

If the antenna length is an exact multiple of quarterwave lengths corresponding to the frequency in use, the antenna will be resonant and will present zero reactance to the transmission line. It should be understood, however, that since the antenna length is constant, the resonant condition will not remain constant for all frequencies. Further, at even multiples of quarter-wavelengths, the resistive component (for end feed) is high, while at odd quarter-wavelength multiples, the resistive component presented by the antenna (at the same feed point) is low.

The function of the antenna tuning equipment is to make the antenna length, together with the selected tuning component, appear at the feed point as some odd multiple of a quarter-wavelength for all frequencies within the tuning range from 0.3 to 26 mc. A block diagram of the antenna coupler and r-f tuner is shown in figure 6-6.

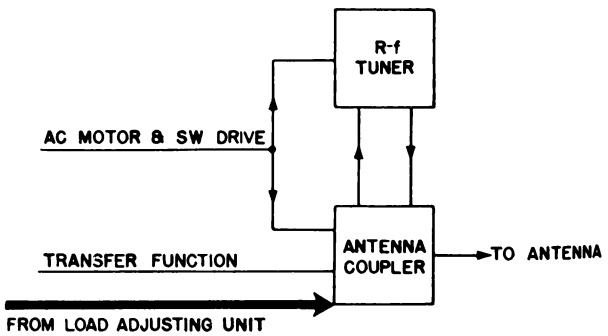

Figure 6-6. –Block diagram of r-f tuner and antenna coupler.

Antenna Coupler

A schematic diagram of the antenna coupler is shown in figure 6-7. The loading switch, S1, permits additional components to be placed into the transmission line to increase its effective length when the length of the main tuning coil is insufficient to achieve the required tuning. This is particularly true at low frequencies when the antenna length should be increased.

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At frequencies where the effective length of an antenna is an even rather than the desired odd multiple of a quarter-wavelength, a capacitor may be switched in the line. Without this action, poorer match would exist between the antenna and the RFA because the antenna impedance would be further separated from the characteristic impedance of the line.

The output of the r-f tuner (discussed later) is permanently connected to the loading switch of the antenna coupler. If one of the loading components is required in the tuning process, the required component is switched into the transmission line through the two-section loading switch, S1. If the loading components are not required, the transmission line is connected directly through the switch shown in the figure. S1 has six positions: one for direct connection and five (ABCDE) for different combinations of loading components. The components selected may be easily located after proper rotation of the switch. The antenna coupler loading control, S2, supplies operating voltage to an actuator, B1, in the antenna coupler. The actuator then functions to position the loading switch as required to select the proper loading component. Table 6-1 shows the components selected by the different positions of the antenna coupler loading control.

Table 6-1.—Components Selected in Each Position of the Loading Switch, S1.

Positions of antenna coupler loading switch, S2	Loading components selected by S1
Direct	None
A	C1 in series
B	C1 in series, C3, C2, shunted across the line
C	L2 in series
D	L1 in series
E	C3, C2, shunted across the line

For example, assume that the antenna coupler loading switch, S2 (fig. 6-7) is placed manually in position A. This action establishes a circuit from one side of the 110-volt a-c line through 11-14 of S2, through contacts NO and C of microswitch S4, to one side of the actuator winding, B1. The other side of the actuator winding is directly connected to the other side of the 110-volt a-c line. The closed contacts NO and C of S4 allow the actuator to rotate until the roller arm of S4 falls into the notch of the positioning cam. This action will open the contacts of S4, breaking the a-c supply to the actuator. The magnetic actuator is ganged to S1 in such a way that the switch is caused to rotate to the A position.

At certain frequencies, the actual antenna impedance is such that no tuning is required to meet the standing-wave ratio reading of 2:1 required at the r-f output of the radio frequency amplifier. At other times, it is necessary to use one or more of the tuning components described in the discussion of the loading switch in the above paragraphs. To meet these conditions, the antenna bypass switch, S3, is introduced into the transmission line (fig. 6-7).

The antenna bypass switch is controlled by the antenna transfer switch on the control indicator (fig. 5-1 of this text). The switch is a double-pole double-throw toggle switch with three positions. These positions are designated TUNER IN, BYPASS, and REMOTE. The REMOTE position is not shown in the figure. S3 is positioned by a magnetic actuator, B2. One side of the a-c line is permanently connected to one side of the actuator. Relay K1, when energized, completes the line voltage circuit allowing the actuator to energize, which, in turn, rotates the positioning cam. In this manner, the antenna bypass switch is caused to position itself in accordance with the position chosen by the antenna transfer switch. Relay K1 receives its operating voltage from the -24 V after time delay line, which is discussed later in this chapter.

With the antenna bypass switch in the BYPASS position (as shown), relay K1 is not energized, and the r-f output from the load-adjusting unit is directly connected to the antenna through J1 and contacts 4 and 1 of S3. When it is desired to have the tuning elements in the line, the antenna transfer switch is placed in the TUNER IN position. S3 is therefore rotated so that the r-f energy now passes through contacts 4 and 3 of S3, and then out through connector J2 to the r-f tuner output. The output of the r-f tuner is permanently connected to the loading switch, S1, discussed in the preceding paragraphs. The loading switch selects the desired tuning component. The output of S1 is connected to contact 2 of bypass switch S3, through contact 1 and to the antenna.

When S3 is switched from TUNER IN to BYPASS, or vice versa, the transmission line is momentarily interrupted, removing the antenna load. When the antenna load is removed, even for

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a short time, it becomes necessary to cut off the radio frequency energy at the IPA (as previously mentioned) to prevent surges in the transmitter output caused by this removal of antenna load. Grounding cam A (fig. 6-8) is fixed to the shaft of the bypass switch, S3, and controls the open or closed condition of microswitch S2. The configuration of the cam is such that it removes the ground from the R9, R10 junction in the IPA control grid circuit during the time that switching takes place. The absence of this ground will cause the IPA to cut off, which, in turn, cuts off the output of the transmitter.

R-F TUNER

The antenna coupler applies its output to the transformer switch, S1 (fig. 6-9) in the r-f tuner. This switch, when connected so that circuits are completed through the 1-2 contacts and the 3-4 contacts, will place the 9:1 impedance transformer in the transmission line. The purpose of the transformer is to step up the signal to the required level. In the other position of the switch (as shown), the impedance transformer is by passed by the connection made between contacts 2 and 4 to the main tuning coil of the r-f tuner.

Magnetic actuator B1, responsible for positioning switch S1, operates under the influence of the transformer switch, S2, through the contacts, NO and C of S3 to one side of the winding of actuator B1. The other side of the actuator is connected directly to the other side of the line. B1 rotates until the roller actuator of S3 falls into the notch of the positioning cam. At

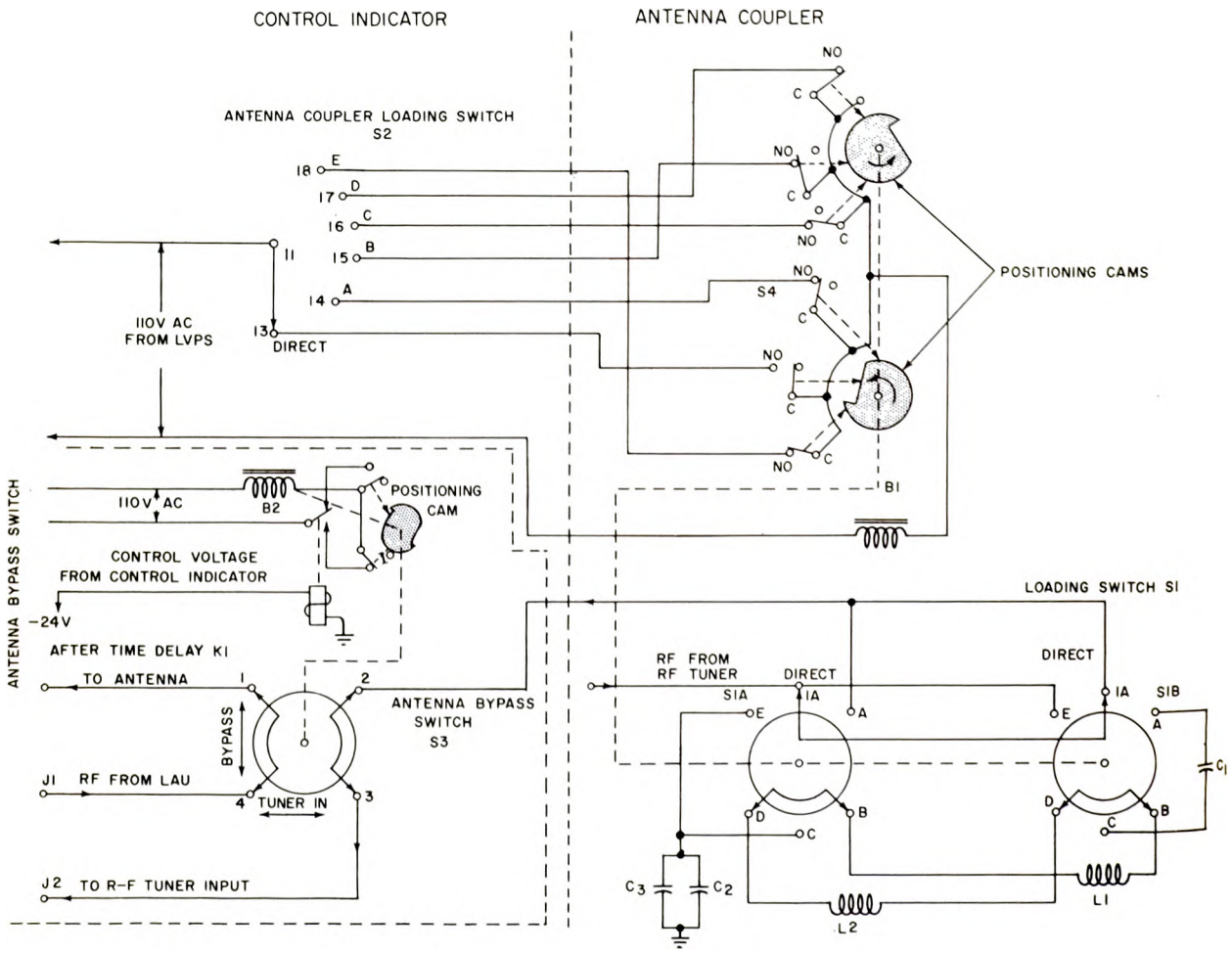

Figure 6-7. –Simplified circuit of antenna coupler.

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this time, the NO and C contacts of S3 open, breaking the line voltage circuit. The transformer switch, S1, is now in the position that eliminates the impedance transformer from the circuit. Placing S2 in position 14 will again energize the circuit through S4. B1 will then cause the positioning cam to rotate until the actuator for S4 falls into the notch of the cam. This corresponds to the position of switch S1, which places the impedance transformer in the circuit.

The transmission line is coupled to the main coil (fig. 6-10) of the r-f tuner by a single loop conductor coil, L1, which is mounted on the shorting ring. The length and configuration of the single loop coil is such that optimum coupling exists above 1 mc if a standard Navy 35-foot whip antenna is used. The resistive component of the antenna impedance is so low at a frequency of 1 mc or lower that the impedance transformer

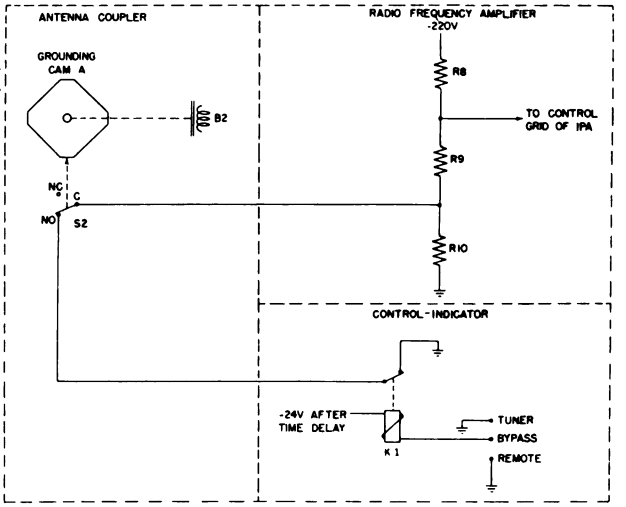

Figure 6-8. —Grounding cam circuit, simplified schematic.

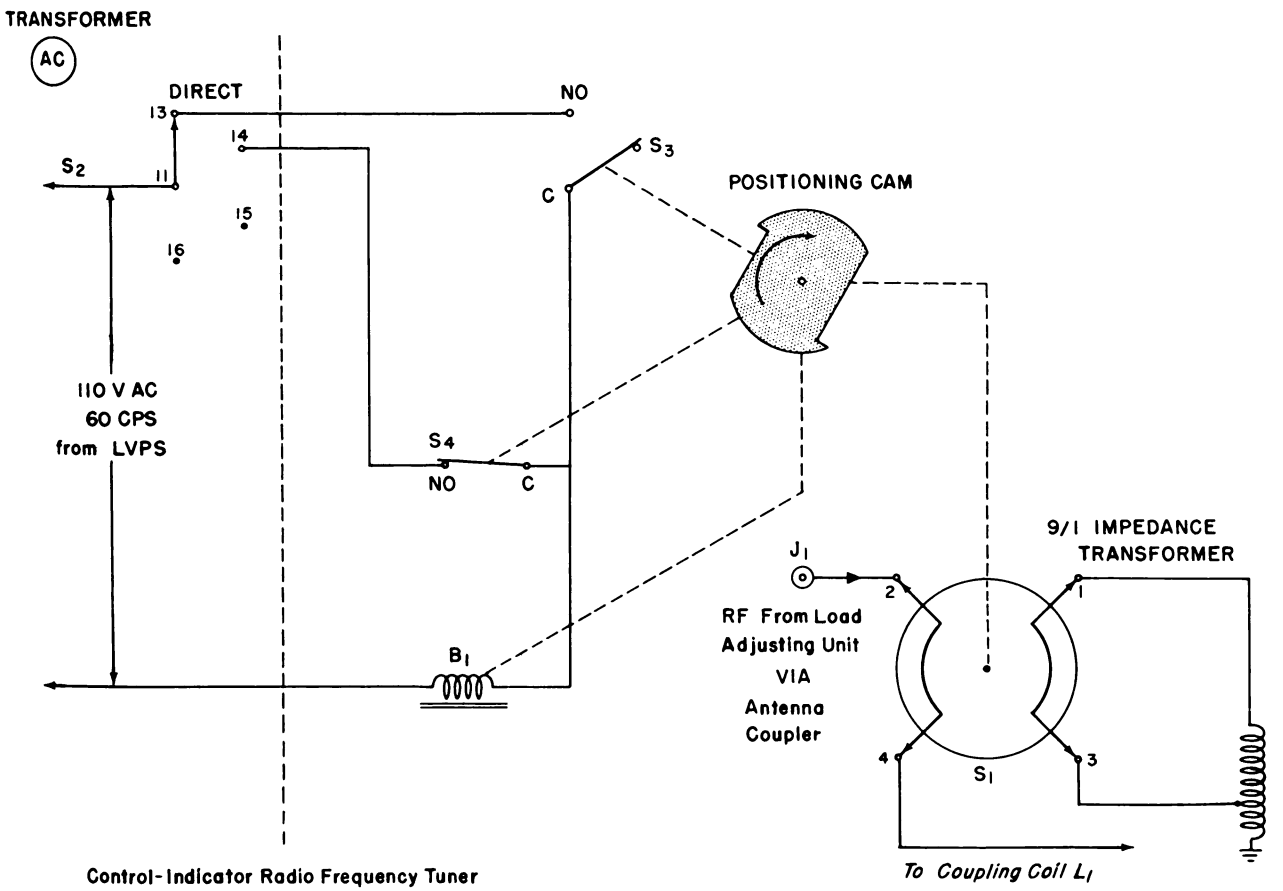

Figure 6-9. —Impedance transformer (r-f tuner), simplified schematic.

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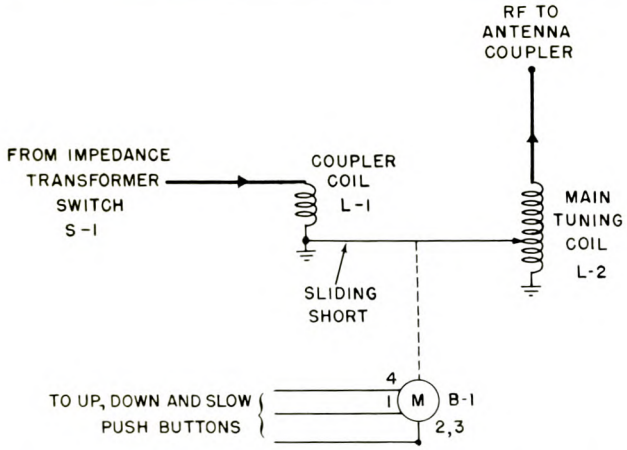

Figure 6-10. –Main tuning coil (r-f tuner), simplified schematic.

will be necessary to bring the radiated signal to the minimum allowable limit.

The main tuning coil of the r-f tuner is equipped with an adjustable sliding short. The tuning coil is the main tuning component in the r-f tuner.

If the impedance transformer (fig. 6-9) is not required in the antenna tuning process, the r-f tuner input to its loading components is connected to the main tuning coil, L2. This tuning coil (fig. 6-10) is a section of a helical center conductor transmission line and is adjustable as forestated, by a sliding short. The short is adjusted so that the effective length of the main coil, together with the effective antenna length, is an odd multiple of a quarter-wavelength at the particular frequency in use. This is the point of resonance or zero reactance.

The position of the sliding short may be adjusted either manually by an adjustment control or remotely by a drive motor, B1. The UP, DOWN, and SLOW pushbuttons control the speed and direction of the motor. The operation of the drive motor is not explained in this chapter. For a detailed discussion of this circuit, refer to Section 2 of NavShips 92121.

ATU BYPASS TRANSFER CIRCUIT

The antenna tuning system may be bypassed in the absence of a transmitter carrier, allowing the antenna to be switched to a receiver. The receiver may be tuned to a different frequency than that of the transmitter. The presence or absence of a carrier is indicated by the condition of the keying line in the low-level radio modulator.

The antenna bypass transfer circuit is mounted in the LLRM for convenience of location only. It might have been mounted in one of the antenna assemblies while receiving its control from the LLRM. The circuitry of the antenna bypass transfer circuit has been taken from the LLRM schematic and is shown in figure 6-11. It is undesirable to have the antenna switching attempt to follow the speed of the hand-key signals. The antenna bypass transfer circuit therefore functions to provide a signal that will place the antenna tuning system in the circuit

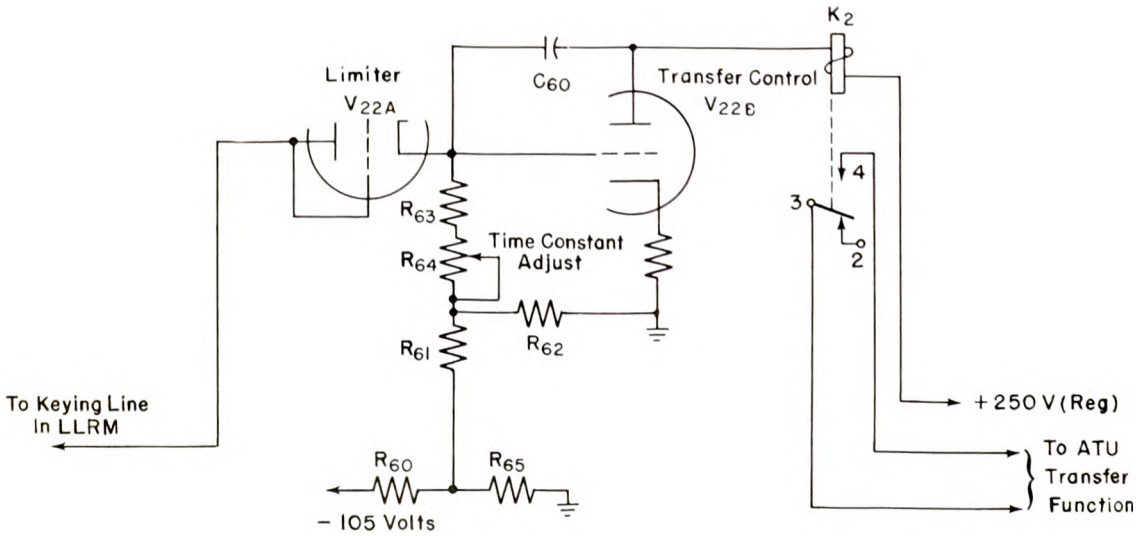

Figure 6-11. –ATU bypass transfer circuit, simplified schematic.

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at the time of the first hand-key signal, and keep the system in contact with the antenna until after the last hand-key signal has been transmitted.

The relay, K2 (fig. 6-11) when energized completes the signal path through contacts 3 and 4 to the antenna tuning system, which keeps the antenna and antenna tuning equipment in the line. Conversely, when deenergized, the contacts open, and the antenna tuning system is bypassed. Reference to figure 6-4 shows that the keying line during key-up condition (S5 contacts open) is at a potential of -24 volts, while in key-down condition (S5 contacts closed), the keying line is at ground potential.

The control grid of the antenna transfer control tube, V22B (fig. 6-11) and the cathode of the limiter, V22A, receive a -10-volt potential with respect to ground (when V22A is nonconducting) through the voltage divider action of R60, R61, and R62 to ground. R61 and R62 are paralleled by R65. The voltage divider is connected from the -105-volt source to ground, and the -10 volts is taken from the junction of R61 and R62.

Since the plate of the limiter section of V22A is connected to the keying line, its potential during key-up condition is -24 volts. With the cathode of V22A at -10 volts and the plate at -24 volts, the tube does not conduct. V22B is also held in the cutoff condition by the -10 volts at its grid. No current therefore passes through the winding of K2, and contacts 3-4 are open. In this condition, the transfer control circuit is open and the antenna tuning system is bypassed. You should note that capacitor C60 connected from the plate of V22B to the cathode of V22A is charged to approximately 260 volts. A charge in excess of the plate supply voltage is evident since the cathode, to which one side is tied, is connected to a point of -10 volts with respect to ground.

On arrival of the first mark (key-down) signal of a hand-key transmission, the keying line is at zero potential, and the limiter, V22A, conducts. The rise in the cathode potential of V22A reduces the grid to cathode bias of V22B, also allowing it to conduct. K2 energizes, closing the antenna transfer control circuit, which, in turn, remove the bypass from the antenna tuning equipment. Since the plate voltage of V22B has fallen and the cathode potential of V22A has risen, the capacitor, C60, discharges to approximately +200 volts.

When the keying signal changes from mark to space (key up) the V22A limiter plate again returns to -24 volts, which renders the tube nonconducting. As V22A begins to move toward cutoff, the cathode potential and, likewise, the grid potential of V22B again moves toward -10 volts. Capacitor C60 begins to recharge. Since the grid of V22B is going more negative, a reduction in plate current is evidenced. The recharge of C60 opposes the tendency of the grid to assume the negative 10 volts and V22B moves slowly toward cutoff. The relay, K2, will remain energized until the current through V22B is reduced to 0.5 milliampere.

The time constant for the charging of C60 and, likewise, the rate at which the grid of V22B returns to the -10 volt potential is determined by the value of K2 winding resistance; the resistance of the B supply; and the value of C60, R63, R64, and R62. R62 is paralleled by R61 and R65. R64 serves as the time constant adjust control, which varies the charge time of C60 from 0.5 to 6 seconds. The relay, K2, can therefore be made to remain energized for any length of time, within this limit, after the final mark signal.

During phone operation, depressing the press-to-talk button produces the same effect on the keying line as depressing the hand-key in hand-key operation. Therefore, the antenna transfer unit bypass transfer circuit functions in the same manner. In machine-key, FSK, and facsimile transmission, a permanent ground is placed on the keying line through the keying equipment, which will keep K2 energized at all times. For this reason, during these modes of transmission, the antenna tuning equipment is permanently in the line.

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QUIZ

1. The radio frequency amplifier (fig. 6-2) accepts from the radio frequency oscillator a signal in what frequency range?
2. The radio frequency amplifier consists of what three stages?
3. (a) The output of the radio frequency oscillator is terminated by what component (fig. 6-2)?
   (b) The final setting of the excitation control (fig. 6-2) should allow an input to the buffer not exceeding what value?
4. The selection of the V1 plate tank circuit components (fig. 6-2) is determined by the setting of what control?
5. After the desired tank circuit of the buffer has been selected (fig. 6-2) it may be tuned to maximum output by varying which component?
6. The r-f output of the IPA (fig. 6-2) is coupled by what component to the control grid of the power amplifier?
7. What two components form a decoupling network for the plate circuit of the PA (Fig. 6-2)?
8. In 100-watt operation of the transmitter, what voltage source supplies the PA screen grid (fig. 6-2)?
9. If K1 (fig. 6-3) is energized, what will be the operating level of the transmitter?
10. (a) The 500 W disable line (fig. 6-5) contains a switch, S2N, which is ganged to what control?
    (b) The switch allows the circuit to be completed in all except what two bands?
11. Energizing antenna switching relay K2 operates what relay in the low-level radio modulator (fig. 6-2)?
12. The load adjusting unit consists of what two circuits?
13. Acceptable tuning of the antenna load by the antenna tuning equipment is achieved if the standing-wave ratio can be reduced to what value?
14. The output of the r-f tuner is permanently connected to what component of the antenna coupler?
15. The antenna bypass switch is controlled by what control on the control indicator?
16. (a) The antenna coupler, applies its output to what component in the r-f tuner (fig. 6-6)?
    (b) What is the purpose of the impedance transformer in the r-f tuner?
17. If the r-f tuner input to its loading components is connected to the main tuning coil, L2 (fig. 6-9) what component is not required in the antenna tuning process?

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CHAPTER 7

OPERATING PRINCIPLES OF A REPRESENTATIVE RADIO TRANSMITTER, PART III

INTRODUCTION

The low-level radio modulator provides the required level of keying voltage to the RFA when the transmitter is operated at low power (100 watts). When the transmitter is operated at the higher power level (500 watts), the low-level radio modulator serves the high-level radio modulator with the necessary input signal amplitude. The high-level radio modulator then provides the required level of output voltage to the RFA. Among the circuits contained in the low-level radio modulator are an audio amplifier chain, a squelch circuit, the automatic gain control circuit, a keying multivibrator, and circuits that control the on-off condition of the transmitter.

The high-level radio modulator is used only when the transmitter is operated at the highest output power. It contains a push-pull power amplifier, which boosts the keying signal to the required level. The high-level radio modulator also contains circuits that control the amount of the plate potential applied to the power amplifier of the RFA. This potential is lower during phone operation to prevent overdriving of the power amplifier during modulation.

The power supplies used with the AN/SRT-14, 15, and 16 are designated low, medium, and high voltage supplies, respectively. Three supplies are necessary because the transmitter at the higher output levels requires higher operating potentials. Several control circuits are contained in the power supplies, which also control the on-off condition of the transmitter. The power supplies are discussed separately at the end of this chapter.

The term, low level, as used in the unit title implies that this particular modulator is used as the modulating source for the RFA during 100-watt operation. It should not be associated with control grid and cathode modulation, or modulation at a point of low r-f potential, which is also referred to occasionally as low-level modulation. Likewise, the term, high level, used in the title of the high-level radio modulator implies that this unit is used during 500-watt operation, and does not refer to the method of modulation.

LOW-LEVEL RADIO MODULATOR (LLRM)

The low-level radio modulator (fig. 7-1) consists of an audio amplifier chain, a keying circuit, a regulated +250 volt power supply, a -12 volt power supply, the antenna transfer unit bypass circuit (discussed in chapter 6 of this training course), and a local remote switch. This unit is commonly called the low-level radio modulator to distinguish it from the high-level radio modulator used in the AN/SRT-15 and AN/SRT-16. A block diagram of the LLRM is shown in figure 7-1.

DISCUSSION OF BLOCK DIAGRAM

The LLRM (fig. 7-1) accepts voice, telegraphy (hand or machine keyed), or facsimile signals at its input. A service selector switch is provided for manual selection of the mode of operation. The unit contains audio amplifying and modulating circuits for the modulation of the carrier at the 100-watt level. As previously stated, during 500-watt operation, the low-level radio modulator feeds a high-level radio modulator, which boosts the audio signal to the required level. A squelch circuit prevents the transmission of noise when no signal is present. Either a carbon or dynamic microphone may be used to apply voice signals to the input.

The electronic keying circuit of the LLRM provides the keying voltage to control the RFA and units 11A, 11B, and 11C of the RFO during amplitude modulation (voice and c-w) and the

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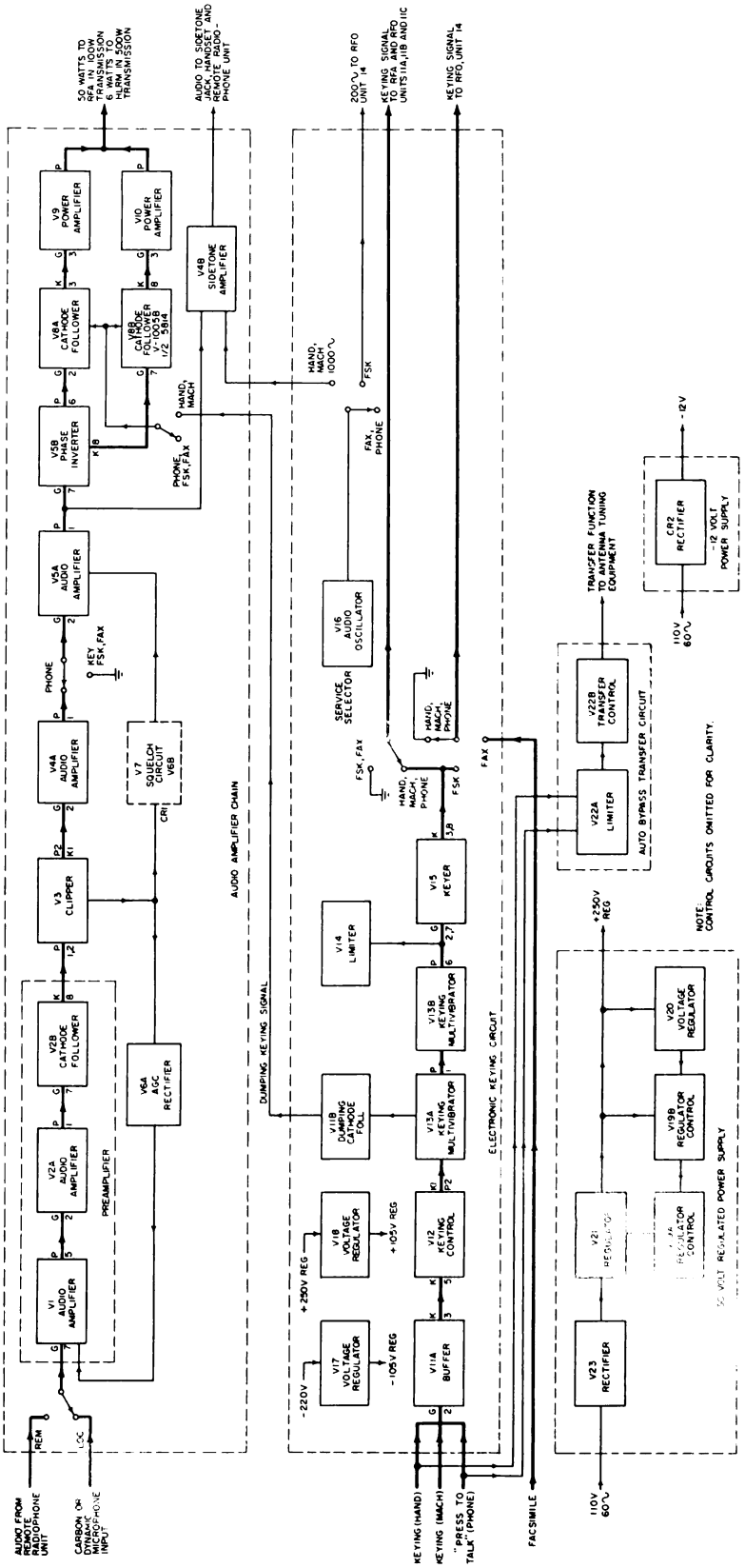

Figure 7-1. -Low-level radio modulator, block diagram.

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radio frequency oscillator only during frequency-shift keying. Facsimile signals also are connected through the low-level radio modulator to the frequency-shift circuits in the radio frequency oscillator.

A regulated +250-volt source is provided to supply the low-level radio modulator and some critical circuits of the radio frequency oscillator. The -12-volt source supplies the microphone.

The front panel of the low-level radio modulator has controls for selecting the mode of transmission, receptacles for a local carbon or dynamic microphone, gain controls, and a squelch circuit control. A test key is provided for carrier frequency control.

CIRCUIT OPERATION

Audio Input

A schematic diagram of the low-level radio modulator is shown in figure 7-2. The audio input signal to the LLRM is received locally through the handset jack, J1, or the microphone jack, J2. The signal is coupled to the 35-ohm tap (terminal 2) of the input transformers, through capacitor C1 and contacts 8R and 10R of the local-remote switch, S1, in the local position. If the input is from a carbon microphone, -12 volts will be supplied to the microphone through contacts 2 and 3 of the dynamic-carbon switch, S2. R1 and R2 and capacitor C2 form a filter network for the -12-volt source. With dynamic microphone input, the -12 volts is disconnected, and the filter network is grounded. When the audio input is applied from a remote position, it is impressed across the full winding of the transformer through contacts 11F and 12F of the local-remote switch, S1. The output of the transformer is connected through a voltage divider comprising R3, R4, and R5 to ground.

To compensate for the lower signal amplitude from the dynamic microphone, the voltage divider output is taken from a higher voltage tap by means of the dynamic-carbon switch, S2. With dynamic phone input, the input to the amplifier, V1, is taken from tap 4 on the divider. When a carbon microphone is used, the input to this stage is taken from tap 6 on the divider. The relative signal levels, as determined by the values of the voltage divider resistors, are such that the following amplifier stages will not be overdriven.

The output from T1 is fed to a two-stage audio preamplifier consisting of V1 and V2A. The output of V2A feeds a cathode follower, V2B, whose function is to match the impedance between the preamplifier stage and the clipper stage, which follows. The cathode follower output is also fed to the automatic gain control (AGC) circuit. The latter circuit provides a negative voltage (under strong signal conditions) to the V1 grid, causing this stage to have a constant output level over a large range of input signal amplitudes.

The audio output from transformer T1 is applied to the suppressor grid of V1 through R6. The AGC voltage is applied to the V1 control grid through R7. The voltage divider network, composed of R8, R9, and R10, supplies the screen and cathode with a positive potential obtained from the +250-volt regulated power source. R10 develops a fixed bias at the cathode of V1 of approximately +2 volts. During no-signal condition, the AGC circuit (to be discussed later) feeds a positive potential with respect to ground to the control grid of V1, causing the V1 grid to draw a small current. In this way the grid-to-cathode voltage is maintained at approximately zero volts. The total suppressor grid to cathode voltage is about -4 volts as a result of the fixed bias and self bias developed from cathode to ground when no signal is received. The tube, V1, operates on the linear portion of its $i_p - e_g$ characteristic curve.

By applying the audio signal input to the suppressor grid of V1 and the AGC voltage to the control grid, less AGC voltage is required to produce a highly sensitive AGC circuit. This represents an improvement over the conventional system of AGC in which both signals are applied to the control grid. In the conventional method, more AGC voltage is required to produce the desired effect.

Cathode follower V2B receives the output variations of amplifier V2A from the plate, which is directly connected to the control grid of V2B. Connecting the circuit in this manner allows a greater amplitude signal to be developed across the cathode follower load resistor, R11. V2B is provided to match the relatively high output impedance of V2A to the low impedance of the diode clipper stage, V3A. The potentiometer, R11, makes possible amplitude control of the signal coupled to the clipper stage. This is identified as the gain to clip control. The entire signal amplitude developed

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across R11 is applied to the squelch circuit and to the AGC circuit.

Automatic Gain Control (AGC)

The signal to the AGC tube, V6A (fig. 7-2) is coupled through capacitor C3. A voltage divider network consisting of R12, R13, and R14, connected from the +250 volts to ground, provides at the junction of R12 and R13 a positive potential of approximately 20 volts. This junction is tied directly to the cathode of the AGC rectifier, V6A, and to resistor R15. If the AGC switch, S3, is set in the ON position (as shown) the potential is further applied through resistors, R16, R17, and R7 to the V1 control grid. Since this potential (with no signal input) is positive, the control grid will draw a current, as mentioned previously, and effectively hold the grid to cathode potential of the tube at zero volts. Because the cathode to ground voltage at no signal is approximately +4 volts, the control grid to ground voltage is also +4 volts. The junction of R15 and R16, at which point the plate of the AGC tube is connected, is approximately +12 volts. Since the cathode is +20 volts at no signal condition, the plate of the AGC tube is negative with respect to the cathode by approximately 8 volts. When a positive peak of the audio signal exceeds 8 volts (with respect to the no-signal voltage) the AGC rectifier, V6A, will conduct, and any additional peak signal voltage will be developed across C3.

During the negative portions of the signal the voltage across C3 is filtered by R16, C5, R17, and C4. The resulting signal, which approaches a d-c voltage is fed through the AGC switch in the ON position to the control grid of V1. The polarity of this voltage across C3 (negative side to the V1 grid) increases the bias on V1, thereby decreasing its gain. The higher the signal appearing across the output of the cathode follower, V2B, the larger will be the negative bias built up across C3. The gain of V1 is controlled by the AGC rectifier so that a 30-db range of input will cause less than a 10-db range of output. The AGC switch may be turned off by placing S3 in the OFF position. In this condition, the V1 control grid-to-cathode potential is maintained at approximately zero volts.

Clipper Stage

Limiting of the audio signal amplitude to prevent overmodulation of the carrier is accomplished in V3 (fig. 7-2). V3 is a twin diode clipper. The B section plate is connected to ground. The A section cathode is +4 volts with respect to ground. This potential is developed between the R13-R14 junction and ground of the voltage divider comprising R12, R13, and R14, which is connected from the +250 volt source to ground. The cathode of the B section, and the plate of the A section, are tied together and held at a d-c potential of between +4 volts and zero volts, depending on the setting of the potentiometer, R14. The potential is made adjustable to allow for differences in the two halves of the input wave-shape to V3. If it is desired to have the two halves exactly similar, the setting of the arm of R14 is placed in the center for +2 volts. With a potentiometer setting at the center of R14, both cathodes are approximately 2 volts positive with respect to their associated plates.

The audio signal from the cathode follower, V2B, is coupled by capacitor C6 to the V3A plate and the V3B cathode. When the positive peak of the audio signal exceeds +2 volts, the V3A plate becomes positive with respect to its cathode, and the A section diode conducts, clipping the positive swing of the input signal to a maximum of 2 volts. Since the tube is effectively an open circuit during the time the cathode is positive with respect to the plate, the audio signal voltage will be developed across R19. When the diode conducts, it is effectively a short across R19, and the unused output is developed during this time, across R18. Likewise, when the negative swing of the audio exceeds -2 volts, the V3B plate will be positive with respect to its cathode, and the B section diode conducts in a similar manner to the A section, holding the maximum negative swing to -2 volts. Potentiometer R14 is designated, clipper symmetry, and may be adjusted so that the positive and negative peaks of the audio signal are clipped at the same level. The clipper output is developed across resistor R19 and is coupled to the audio amplifier, V4A.

Audio Amplifier

V4A (fig. 7-2) receives the output of the clipper through capacitor C7, amplifies the signal, and couples this signal through capacitor

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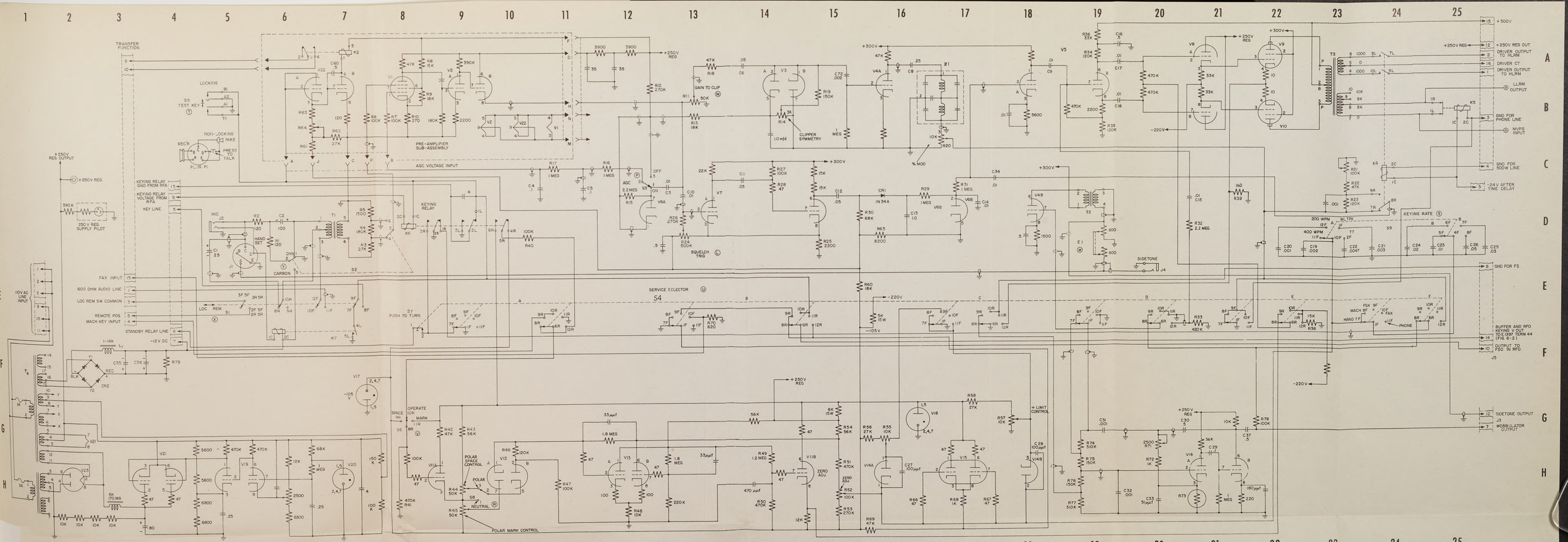

Figure 7-2. –Low-level radio modulator, simplified schematic.

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zero volts.

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C8 to the bandpass filter network, Z1. The filter is designed to pass frequencies in the range from 200 to 4500 cycles and assist in cutting out high-frequency harmonics of the audio signal introduced by the clipper action in the previous stage. The circuit also eliminates noise frequencies outside the pass band of the filter. The output of the filter is developed across potentiometer R20, which is known as the % modulation control.

In 100-watt operation, the low-level radio modulator delivers 50 watts to the radio frequency amplifier to modulate the carrier, and during 500-watt operation it delivers 6 watts to the high-level radio modulator, which, in turn, modulates the carrier. For this reason, the output of potentiometer R20 is developed across a voltage divider under the control of the gain change relay, K4. The voltage divider comprises R21, R22, and R23. During 100-watt operation the entire output is coupled to the V5A stage through contacts 7R and 8R of the unenergized relay, K4. In 500-watt operation, K4 is energized, and the output of R20 is developed across the voltage divider and tapped from the junction of R22 and R23. The signal is then fed through relay contacts 8R and 9R of K4 to the input of V5A. Depending on the setting of R21, the audio signal fed to the amplifier, V5A, in 500-watt operation is 30 to 50 percent of the signal output from R20. The output of R20 is adjusted for 100-percent modulation of the carrier in either 100- or 5-watt operation.

With the service selector switch, S4, in any position other than phone (switch positions are indicated on S4F), the output of V4A will be grounded through S4E front. The service selection switch, S4, is divided into front and rear sections. The front section is designated F on the switch contacts, and the rear section uses the letter, R. In phone operation, the keying relay, K6, controls the ground on potentiometer R20 in the following manner. When the relay is not energized, the circuit is from the arm of R20 through S4F (contacts 12R and 6R) to relay K6 (contacts 1R and 2R) and to ground. To energize relay K6 depress the press-to-talk button. This action described later in more detail removes the ground on R20 by opening contacts 1R-2R of K6 on the handset. The signal then passes to the input of V5A through capacitor C34 and either the 7R-8R contacts or the 8R-9R contacts of K4.

In addition to the audio signal being coupled to the grid of V5A, there is a negative bias applied from the squelch circuit, V7, and V6B (discussed in the next paragraph), which is sufficient to cut off V5A if the level of the audio is not high enough. The output of V5A is coupled through C9 to the phase inverter, V5B, and to the sidetone amplifier.

Squelch and Squelch Rectifier Circuits

A squelch circuit is provided to prevent the transmission of noise in the absence of an audio input signal (when the press-to-talk button is depressed). The squelch circuit prevents amplification by biasing the V5A grid beyond cutoff. The squelch circuit consist of a one-shot multivibrator, V7, a crystal rectifier, CR1, and a limiter V6B (fig. 7-2).

The audio signal as developed at the output of the cathode follower, V2B, is coupled through capacitor C10 to the grid of V7A. Resistors R60, R24, and R25 form a voltage divider network from -105 volts to ground. (R65 is in parallel with the latter two resistors to ground.) This produces a potential of -30 volts (with respect to ground) at the junction of R60 and R24. R24 is the squelch trigger control and has its arm tied to the grid of V7A through grid leak resistor R26. With the control turned all the way (left), the grid of V7A would normally be -30 volts to ground, but V7B, whose grid is tied to +300 volts through R28 and R27, conducts heavily. Its conduction causes a positive voltage to appear across the common cathode resistor, R25, of approximately 20 volts. The -30 volts from the voltage divider adds to the 20-volt drop across R25 to hold V7A well below cutoff. Turning the squelch trigger control in the other direction reduces the bias on V7A.

The audio signal coupled to the grid of V7A must be large enough to overcome the grid-to-cathode bias to effect the operation of the squelch multivibrator. When V7A conducts, its plate voltage drops. The decrease in plate voltage is coupled through C11 to the grid of V7B and causes the plate current in this section to decrease. The decrease in plate current of V7B also reduces the bias voltage developed across R25, which allows V7A plate current to increase more. The action is rapidly accumulated until V7B is cut off. At the end of the discharge of C11, V7B again conducts and remains in the conducting state until the next positive pulse of the audio signal appears on the V7A grid. The output pulses of the multivibrator are coupled to the squelch rectifier, CR1.

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Squelch rectifier CR1 and squelch limiter V6B receive the positive pulses from the multivibrator by way of C12. R60 and R65 form a voltage divider from the -105 volts to ground to establish a -30 volts potential at the R60-R65 junction with respect to ground. When there are no positive pulses from the squelch multivibrator, the -30-volt potential appears at the grid of the audio amplifier V5A. This potential biases V5A to cutoff. The -30 volts is applied through R30, the high back (leakage) resistance of the crystal rectifier, CR1, R29, and R31. When there are positive pulses from the multivibrator, they appear across R30, and CR1 conducts. As C13 discharges through CR1 the C13 voltage is reduced. The output of the filter is applied to the V5A grid. This action raises the grid potential of V5A above cutoff, allowing it to amplify the audio signals being received from V4A.

The time constant of the filter circuit is sufficiently long to allow the bias voltage to be essentially constant and at a level needed to keep the tube operating in the linear portion of its characteristic curve.

The diode limiter, V6B, acts to prevent the application of a positive potential to the V5A grid. If the positive pulses from the multivibrator are of high enough amplitude to make the resultant voltage applied to V5A grid positive with respect to ground, the plate of V6B becomes more positive than its cathode. The diode therefore conducts heavily, limiting the voltage applied to the grid of V5B to a maximum of zero volts with respect to ground. Peak positive voltages above the ground level are developed across resistor R29.

The Sidetone Amplifier

The output of the audio amplifier, V5A, is monitored by the sidetone amplifier, V4B. The signal is coupled by C15 to a voltage divider comprising R32 and R33. A reduced portion of the signal, taken from the junction of the two resistors, is connected through contacts 12R and 6R of S4D (when the service selector switch is in the phone position) and applied to the grid of the sidetone amplifier, V4B.

The output of the sidetone amplifier, V4B, is coupled by an output transformer, T2, to three monitoring points: the handset receptacle, J1, the sidetone jack, J4, and the remote radiophone unit via J3. The output of J1 and J4 is under control of the sidetone attenuator, E1, which is the audio volume control. In addition, the L-pad, E1, acts as an impedance matching network, which presents a constant impedance to the output transformer, T2, regardless of the setting of E1. The full sidetone output is applied to the radiophone unit since it is equipped with a separate L-pad (not shown).

A 1000-cps signal from the audio oscillator, V16, is applied to the grid of V4B in hand-key or machine-key operation during the key down (mark) condition for monitoring purposes. Switch S4D connects the signal through its contacts, 8R and 6R, in hand-key operation, and through contacts 9R and 6R in machine-key operation.

Audio Output Circuit

The output signals of V5A are fed to the phase inverter, V5B, which produces two signals of equal amplitude, but of opposite phase. These two signals are coupled to a push-pull cathode follower, V8, which acts as a driver for the push pull power amplifiers, V9, that constitute the final stage of amplification of the audio amplifier chain. In the AN/SRT-14, the final audio output (50 watts) is fed directly to the radio frequency amplifier. It should be recalled that the AN/SRT-15 and 16 provide two levels of audio output, depending on the power level at which the transmitter is operating. For 500-watt operation, the output (6 watts) of the LLRM is fed to the HLRM.

Phase Inverter

The circuit components of the phase inverter, V5B, are such that two waveforms, equal in amplitude but opposite in phase, will be developed at the output. The waveforms are taken from the cathode and plate, respectively. The voltage across the 2200-ohm resistor in the cathode circuit provides the operating bias. The degenerative action across this resistor is such that the output waveforms are of the same amplitude as the input signal. The plate resistor, R34, and the cathode resistor, R35, are of equal value to ensure equal output signals from cathode and plate. R36 and C16 decouple the stage from the +500-volt power supply.

Push-Pull Cathode Followers

V8A and V8B (push-pull cathode followers) act as drivers for the push-pull power amplifiers, V9 and V10. Since the power amplifier

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stage is operated class AB2, the cathode followers V8A and V8B are required to supply energy to a low input impedance during the portion of the cycle when grid current flows in V9 and V10.

With the service selector switch in the phone position the V8 grids are returned to a d-c potential of -40 volts by way of contacts 6R and 12R of S4E (rear). This potential is obtained from the voltage divider comprising R37 and R38, which is connected from the -220 volt supply to ground. The V8 cathodes are returned to the -200-volt source, allowing the tubes to conduct during no-signal condition. The V8 cathode potential developed as a result of this current flow is approximately -40 volts.

The control grids of the push-pull power amplifier tubes, V9 and V10, are maintained at the same potential as the V8 cathodes through direct connections. Signals of equal amplitude but opposite phase (from the phase inverter) are coupled to the grids of V8A and V8B through capacitors C17 and C18, respectively. The control grids of the power amplifiers are held at -40 volts with respect to ground during no signal condition. When the signal from V8 exceeds 40 volts the control grids of the power amplifiers, V9 and V10, will become successively positive with respect to their cathodes, and grid current will flow. When the power amplifier grids conduct, they present an impedance of approximately 1000 ohms to the cathode followers. For example, if the signal to the power amplifier is 50 volts peak, the amplifier grids will successively become 10 volts positive with respect to their associated cathodes. The cathode follower, V8, is capable of conducting this current without clipping because of the relatively low impedance of the V8 cathode circuit. At the same time, the high input impedance of the V8 grid circuit is maintained since the V8 grids do not go positive with respect to their associated cathodes. Thus the required driving energy can be supplied to the power amplifier while maintaining the high input impedance to the cathode followers, V8A and V8B.

The audio amplifier chain is not required to amplify audio signals during hand-key and machine-key operation. Therefore, the cathode followers and power amplifiers perform a different function during these two modes of operation. The fact that the power amplifiers use +500 volts for a plate supply is utilized to provide a dumping (loading) circuit to eliminate transient voltage variations imposed on the power source when the transmitter is keyed. The dumping control is fed to the grids of the cathode followers, V8A and V8B, through the service selector switch, S4E (rear), when in either the HAND or MACHINE position.

The +500-volt power supply provides plate voltage for the IPA and the screen voltage (during 500-watt operation) to the PA in the radio frequency amplifier. In space condition the RFA is keyed off, and the IPA plate current and PA screen current are at minimum, while in mark, they are at maximum. If these sudden changes in load current drawn from the 500-volt supply are not modified, transient voltages will appear across the power supply filter chokes.

At the start of a mark condition, the polarity of the transient voltage in the power supply filter will be such as to oppose the sudden rise in current, thus causing a dip to appear in the power supply output voltage. Conversely, at the start of a space condition, the transient voltage will be aiding, opposing the sudden decrease in current, thus increasing the supply output voltage.

The function of the dumping circuit is to provide a load during space condition only, to compensate for the reduction in load current when the IPA is keyed off. Thus, the load to the 500-volt supply will be approximately equal during mark and space due to this elimination of transient variations. The method employed to keep the power supply current from decreasing during a space is to supply a keying voltage from the keying circuit of -65 volts during mark and zero volts during space to the grids of the push-pull cathode followers. This voltage is obtained from the cathode of the dumping cathode follower, V11B. (Note that the cathode returns to -120 v.) The input to the cathode followers is reproduced at the cathodes, which are directly tied to the grids of the power amplifiers. During space, with the grids at zero, the power amplifiers conduct, while during mark, the amplifiers are biased below cutoff by the -65 volts. During space, when the tubes conduct, the voltage developed across the common cathode resistor, R39, acts as a degenerative bias so that the two tubes draw 120 ma total from the +500-volt supply. This value of current is equivalent to the reduction caused by a space keying of the radio frequency amplifier.

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Power Amplifiers

As previously stated, the no-signal grid bias of the power amplifiers, V9 and V10, applied by way of the V8 cathode followers is approximately -40 volts. With a screen voltage of +300 volts and a plate supply of +500 volts, the value of bias is such that the stage operates as a class AB2 push-pull amplifier. Operation at this level allows high-power outputs with high plate efficiency and low distortion. With the service selector switch, S4C (rear), in the phone position, the common cathode resistor, R39, of the power amplifiers is shorted to ground through its contacts 6 and 12. In the HAND and MACH positions, R39 is in the circuit, adding degeneration so that the push-pull amplifiers operate at the proper current drain values required for the dumping action as discussed previously.

The modulation transformer, T3, is provided with two secondaries. The 50-watt secondary is used only during 100-watt operation of the transmitter, and the 6-watt secondary is used during 500-watt operation. The output of the 6-watt secondary is connected through contacts 7L, 8L and 9L, 10L of the energized relay, K4, to the grids of the amplifiers in the high-level radio modulator. Negative bias of 50 volts for the high-level radio modulator is applied through the center tap of the 6-watt secondary. K4 energizes from -24 V after time delay to the GROUND FOR 500 W AX line. (See figure 6-3, chapter 6.) The AN/SRT-14 operates at the 100-watt level only and is not controlled by the 500 W disable line, since this line is nonexistent because of the absence of the high-voltage power supply. With the AN/SRT-15 and 16 in 100-watt operation, K4 is deenergized, and the circuit of the 6-watt secondary is opened. The AN/SRT-14 also contains this relay which is deenergized at all times.

If relay K5 is deenergized, the 50-watt secondary of T3 is short-circuited by contacts 1L and 1R of K5, preventing voice modulation. The plate voltage is delivered through these contacts to the final amplifier in the RFA. In the AN/SRT-14 the plate voltage is delivered directly to the RFA through a switch (discussed later) in the HLRM. With the service selector switch, S4E, in the phone position, K5 is operated from ground through contacts 1 and 11 of S4E of the -24 V after time delay. This source of ground is known as the GROUND FOR PHONE line and is also utilized in the control circuits of the RFA, MVPS, and HVPS. With K5 energized, the short is removed from the 50-watt secondary, and the plate supply voltage to the RFA is amplitude modulated by the 50-watt audio signal.

KEYING CIRCUIT

In hand-key operation the keying circuit accepts keying signals that determine the on-off condition of the carrier. In machine key, keying signals are sent to the keying circuit from a teletypewriter or other keying equipment, which also control the on-off condition of the carrier. In both hand-key and machine-key operation an audio oscillator, V16, produces an audio signal for aural monitoring of the keying signal. In FSK operation, keying signals are received by the keying circuit from machine-key equipment, and signals are produced that control the operation of the frequency-shift oscillator in the RFO. Signals are received from an audio oscillator, which phase-modulate the carrier to minimize the effects of selective fading. In phone operation, audio signals are fed to the keying circuits via the press-to-talk button on the microphone, which also keys the carrier on and off.

The keying circuit of the LLRM consists of the keying relay, K6; press-to-talk relay, K7; buffer, V11A; keying control, V12; keying multivibrator, V13; dumping cathode follower, V11B; limiter, V14; keyer, V15, and audio oscillator, V16. The outputs of the keying circuit and their functions are listed in table 7-1.

The keying input circuit, containing service selector switch, S4A, selects the input to the keying circuit. In hand-key and phone operation, the keying signals are introduced to the LLRM through the action of the keying relay, K6 (refer to figure 7-3). With the local-remote switch, S1, in the REM position, the keying line is completed to ground in a radiophone unit through contacts 8F and 9F of S1 (if the key is in the MARK or DOWN position). When the key is up, there is no ground. Therefore, the transmitter is keyed on by a mark, and off by space in c-w operation. The antenna switching relay, K2, is also operated to switch the antenna to the transmitter when the ground is supplied. When in phone operation, the press-to-talk button also supplies ground to the keying line when it is depressed.

If the phone is connected locally, S1 is in the LOCAL position, and the ground from the press-to-talk button operates K7 from the -12-volt supply. With K7 operated, ground is fed to

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Table 7-1. —Functions of Keying Circuits Outputs.

Output	Delivered to	Function
RF keying signal	Buffer in RFA and output of RFO	To key the carrier
FSK keying signal	FS oscillator in RFO	Control frequency of FSO
Keying signals to dumping circuit	LLRM	Load compensation for +500-volt power supply
200-cycle signal	FS oscillator in RFO	Phase-modulate the FSO
1000-cycle signal	Sidetone amplifier in LLRM	Monitoring of keying signals

K6 through contacts 5L and 4L of K7 and contacts 7F and 9F of S1.

The test key, S5, parallels the keying line so that depressing S5 also supplies a ground to operate K6. With the service selector switch, S4, in either FSK or FAX position and the local remote switch, S1, in the REM position, a permanent ground is placed on the keying line through contacts 10R and 6R or 11R and 6R of S4B, from the keying equipment.

Two gas tube voltage regulators, V17 and V18 (fig. 7-2), supply a regulated -105 volts and a regulated +105, respectively, for use in the keying circuit. The -105 volts is connected to contact 4R of relay K6, and the +105 volts is connected to contact 6R. In hand-key operation, when the key is depressed (mark), K6 operates, and +105 volts is connected through contacts 6R and 5R of K6, resistor R40, contacts 8R and 6R of S4A, contacts 8R and 10R of S6 in the operate position, and resistor R41 to ground. When the key is up (space), K6 is not energized, and -105 volts is now applied through contacts 4R and 5R of K6 and through the same circuit as stated above. R40 and R41, acting as a voltage divider across either V18 or V17, provide either +85 volts or -85 volts for mark and space, respectively, to the grid of the buffer stage, V11A, during hand-key operation. In phone operation, the same control of the buffer stage is accomplished through contacts 12R and 6R of S4A, except that the press-to-talk button replaces the hand key. The result of this action applies +85 volts to the grid of the buffer when the press-to-talk button is depressed and -85 volts to the grid when the button is released.

A permanent mark or space signal can be produced at the grid of the buffer by the action of S6. Setting this switch to mark connects +105 volts to the grid of the buffer through contacts 11R and 8R of S6. In the space position, S6 connects -105 volts through its contacts 8R and 9R to the buffer grid.

In addition to providing keying voltages for the keying circuit, the keying relay, K6, also controls the action of the standby relay, K1, located in the low-voltage power supply shown in figure 7-6. The operation of K1 provides the -24 V after time delay, which controls the operation of the MVPS, HVPS, and HLRM. The latter two units are found in the AN/SRT-15 and 16 only. The ground to operate K1 is furnished by way of the push-to-turn switch, S7 (fig. 7-2) and contacts 3R and 2R of energized K6. In hand-key operation only, the action of K6 in controlling K1 (LVPS) is bypassed by a permanent ground supplied for the standby relay line through contacts 1F and 7F of the service selector switch, S4A. The standby relay, K1, would be unable to follow the keying signals, making it necessary to keep the relay energized at all times during hand-key operation. In other modes of transmission, the transmitter cannot go on the air unless K6 is energized, supplying the ground to operate K1. In machine-key, FSK, and facsimile transmission, a ground is supplied over the keying line from the remote radiophone unit that energizes K6 and provides a standby operate control for K1. In phone operation, K6 is operated only when the press-to-talk button is depressed, either locally or remotely. Therefore, the output circuits of the power supplies controlled by K1, which, in turn, is controlled by K6, are completed in phone operation only when the press-to-talk button is depressed.

Buffer

The purpose of cathode follower V11A (fig. 7-2) is to isolate the action of the keying input signals from the keying multivibrator, V13. R42 in the V11A plate circuit is a parasitic suppressor. The cathode of V11A is power biased to +40 volts by the voltage divider comprising R43, R44, and R45, which is connected from the regulated +105 volts to the regulated -105 volts. The keying voltages for the different modes of operation are applied at the grid of

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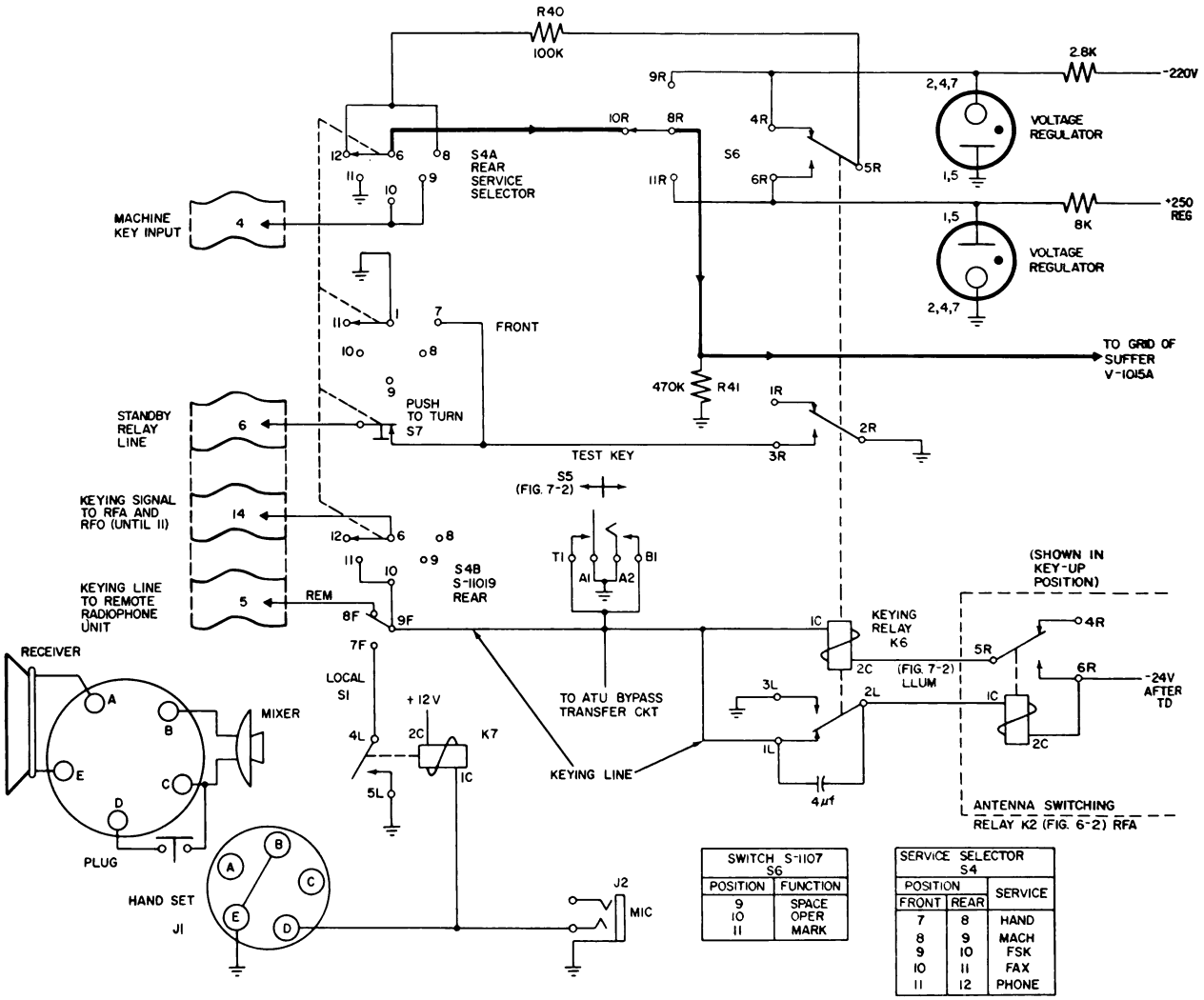

SWITCH S-1107 S6	
POSITION	FUNCTION
9	SPACE
10	OPER
11	MARK

SERVICE SELECTOR S4		
POSITION		SERVICE
FRONT	REAR	
7	8	HAND
8	9	MACH
9	10	FSK
10	11	FAX
11	12	PHONE

Figure 7-3. –Keying input circuit, simplified schematic.

V11A, and are reproduced at the cathode. When the negative input exceeds approximately -45 volts, V11A will cut off, with the result that the maximum amplitude of a negative keying signal appearing at the cathode is about 40 volts.

Keying Multivibrator

The keying control tube, V12 (fig. 7-2) and keying multivibrator, V13A, receive the output of the cathode follower. In the space condition V13B conducts, causing current to flow through the common cathode resistor, R48. The voltage developed across this resistor is sufficient to cut off V13A even though a small positive voltage is developed between the V13A grid and ground. The keying control tube, V12B, assures that V13B conducts and V13A is cut off during space. The plate of V12B is connected to the junction of R46 and R47 located in the grid circuit of V13A. If V12B were not included in the circuit, the junction of the resistors would be about +5 volts during space.

The cathode of V12B during neutral keying is tied to the cathode of the buffer, V11A, through contacts 2 and 1 of the neutral polar switch, S8, in the NEUTRAL position. During space this point is at either +4 volts for neutral machine keying or -40 volts for hand keying. As these values are at or below the normal +5 volts to which the plate of V12B is tied, V12B conducts, effectively tying the junction of R46 and R47 to

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the space voltage appearing at the buffer cathode. Therefore, the grid circuit of V13A during space is at a lower (less positive) potential than the grid of V13B, causing the desired condition of having V13A cut off and V13B conducting. When a spacing signal is applied the same action occurs in polar keying, except that the keying control, V12B, is connected so that its cathode goes to about +4 volts when the grid of the buffer, V11A, is at -30 volts. This action is accomplished by connecting the cathode of V12B to a tap on potentiometer R44 through contacts 2 and 3 of S8 in the POLAR position.

R44 is called the polar space control and is a part of a voltage divider consisting of R43, R44, and R45, which is connected between the +105- and -105-volt supplies. The tap of the potentiometer is tied to the cathode of the buffer. In space, with a -30-volt keying signal on the grid of the buffer its cathode will be a potential of about -27 volts. The setting of R44 will be such that the tap will be about +4 volts. With the multivibrator in the space condition, the plate voltage of V13A will be approximately +235 volts, and the plate of V13B approximately +100 volts.

As the keying signal changes from space to mark, the potential of the buffer cathode starts to rise. As this potential reaches +5 volts in neutral keying or -27 volts in polar keying, the keying control tube, V12B, no longer conducts and has no effect on the grid of V13A. The potential at the latter grid rises as a result of this action, but does not rise sufficiently high to overcome the common cathode bias developed across R48. As the mark potential rises higher keying control tube, V12A, comes into action. The cathode of V12A is tied to the grid of V13B. The plate of V12A is connected to the tap of potentiometer, R45, which is the cathode resistor of the buffer circuit. One end of the buffer cathode resistor is tied to the -105 volt source.

The potentiometer, R45, is known as the neutral-polar-mark control. In order for V12A to conduct, its plate potential must be higher than that at the grid of V13A. As the mark voltage increases at the cathode of the buffer, the voltage of the tap of R45 will become higher than the grid of V13A, allowing V12A to conduct. When V12A conducts, it effectively makes the potential at the grid of V13A the same as that developed at the tap of R45. The tap is set so that as the mark signal reaches +30 volts at the grid of the buffer, the voltage at the tap, and also at the grid of V13A, will be sufficient to overcome the bias potential developed across the common cathode resistor, causing V13A to conduct. The tube remains in conduction throughout the marking impulse. Therefore, in the mark condition, the plate of V13A is at a potential of approximately +70 volts, while the plate of V13B has a potential of about +235 volts.

As the keying voltage changes from mark to space, the potential of the grid of V13A must be lowered enough to cause a reduction in the tubes current flow, which will initiate the cumulative flip-flop action. During the mark condition, the junction of R46 and R47 (to which the plate of V12B is tied) is about +11 volts. Therefore, when the keying signal reduces to a point where the cathode of V12B (which is tied to the cathode of the buffer in neutral keying and to the polar space control in polar keying) goes below +11 volts, V12B conducts, lowering the grid potential of V13A. In neutral keying, when the keying signal has reduced to about +7 volts, V12B will conduct a sufficient amount to lower the V13A grid below the saturation point, and the flip-flop action takes place. In polar keying the polar space control, R44, is set so that when the negative or space keying signal reaches about -28 volts, the keying control, V12B, conducts enough to lower the grid potential of V13A below the saturation value, and the multivibrator returns to the space condition.

The keying multivibrator, V13, has two outputs. One of the output voltages is developed in the plate circuit of V13A, and the other, which is out of phase with reference to the first, is developed in the plate circuit of V13B. The keying voltage developed in the plate of V13A is fed to the dumping cathode follower, V11B, for use in the push-pull cathode followers of the audio amplifier chain of the LLRM. A keying signal of zero volts for space and -65 volts for a mark is delivered to V11B from a voltage divided in the plate circuit of V13A. The divider comprises resistors R49 and R50, which are connected to the regulated -105 volts. The dumping cathode follower provides a low impedance to the input of the push-pull cathode followers. The time constant in the output circuit of the dumping cathode follower is such that the waveshape is not destroyed. The output is taken directly from the cathode of V11B. The output signals are connected to the push-pull cathode followers through the service selector switch, S4E (rear), in the HAND and MACHINE positions. In frequency-shift keying and facsimile operation, the transmitter is functioning continuously, and

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no load compensation is required. In phone operation, load compensation is not required since the +500-volt supply is turned off when the press-to-talk button is up by the action of the keying relay, K6.

The keying signal from V13B, which is +25 volts for mark and -30 volts for space, is delivered to the limiter, V14A and V14B. This range of voltage is obtained from a voltage divider connected from the plate of V13B, through R51, zero adjust control R52, and R53, to the regulated -105 volts. R54 connects the voltage divider to the +250 regulated supply. The keying signal is developed across a value of capacitance that is selected by the keying rate switch, S9 (center right of schematic). Either of the capacitors C19, C20, C21 or C22, may be chosen to lengthen the circuit time constant as required. The time constant involved in charging and discharging this capacitance should give the required rise and decay time to the keying signal waveshape. In HAND and MACH positions of the service selector switch, a second set of waveshaping capacitors comprised of C23, C24, C25, and C26 is connected in parallel with the first set. The parallel capacitors are connected through S4F (contacts 7F and 1F for hand and contacts 8F and 1F for mach operation).

Limiter

The plate of the limiter, V14A, is tied to the tap of a potentiometer, R55. R55, known as the limit keyer out control, is part of a voltage divider network from the regulated -105 volts through R56, R55, and contacts 9F and 1F of service selector switch S4C to ground. R55 is adjusted for -10 volts at the tap. Therefore, any negative (space) keying signals in FSK operation are clipped at a maximum of 10 volts by the action of the limiter diode, V14A. In other modes of operation the ground is removed from R55 by the service selector switch, S4C, and the plate of V14A is tied to the -105 volts.

In hand-key, machine-key, and phone operation, the cathode of the limiter, V14B, is grounded through service selector switch, S4C, through contacts 7F and 1F for hand, 8F and 1F for mach, and 11F and 1F for phone. Note that in these positions of the switch, potentiometer R57 in the cathode of V14B is shorted through the switch contacts. In FSK operation, the short on the cathode of V14B is removed. The cathode is then connected to the tap of the potentiometer R57, which is part of a voltage divider from +105 volts, through R58, R57 to ground. R57 is the (+) limit control, and is adjusted for +10 volts at the tap. In this condition, V14B clips positive (mark) keying signals at +10 volts. Therefore, positive (mark) signals in hand-key machine-key, and phone operation are clipped at zero, whereas in FSK operation, they are clipped at +10 volts. C27 and C28 are bypass capacitors for r-f energy.

Keyer

The keying signals, as supplied by the limiters, are fed to the grids of the keyer, V15. This stage utilizes a twin triode operating in parallel, with parasitic suppressors R66 and R67 connected in the respective control grids. The cathode load is made up of R68 and R69, which constitute a voltage divider with the output taken from the junction of the two. When a mark keying signal (zero volts) is applied to the grid of the keyer, the cathode is driven to about 2 volts positive. Since the stage is operated class A, the mark signal as applied to the buffer of the RFA should not go above zero volts. For this reason the output is taken at the junction of R68 and R69 to reduce the mark voltage from +2 volts to zero.

The output of zero volts for mark and -30 volts for space is the same in hand key, machine key, and phone operation. These signals are used to cut off the buffer in the RFA and the output stage of the RFO in the KEY UP (space) condition for the two keying modes or the press-to-talk button UP in phone operation. Both the buffer in the RFA and the output of the RFO are keyed off to assure that the key-up radiation from the transmitter group is below the minimum allowable level of 400 $\mu\mu$w.

In FSK operation the output from the keyer, which is +10 volts for mark and -10 volts for space, is fed to the radio frequency oscillator to frequency modulate the frequency-shift oscillator in the RFO. Contacts 4R and 6R control the keying circuits that supply the FSK output to the FSO in the RFO. The signals are fed from the junction of R68 and R69 through the shorted contacts 8R and 9R of S4B (rear), through contacts 9F and 1F of S4B (front), to terminal 10 on J5. Facsimile signals are developed across resistor R70 and fed through contacts 10F and 1F of S4B (front), to the output terminals. The other positions of the switch S4B (front) are grounded since the frequency-shift oscillator is

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not used in hand-key, machine-key, or phone operation.

Audio Oscillator

A Wien bridge circuit comprising V16 (fig. 7-4) functions as the audio oscillator and provides two outputs. In FSK operation the output is a 200-cycle signal whose purpose is to phase modulate the frequency-shift oscillator in the RFO whenever it is necessary in order to prevent the effects of selective fading. The amplitude of the 200-cycle signal is sufficient to produce a one-radian phase shift in the carrier. This amount of shift can be varied between zero and one-radian by a PHASE MOD control (not shown). In hand-key and machine operation, the audio oscillator produces a 1,000-cycle signal during the key down (mark) condition that is fed to the sidetone amplifier in the audio chain for monitor purposes.

In order to sustain oscillation, a feed-back signal must be developed that is large enough to overcome the losses in the grid circuit, and must be in phase with the signal on the grid. Signals appearing at the plate of V16A are coupled through C29 to the grid of V16B, which is an amplifier and phase inverter. The signals appearing at the plate of V16B are in phase with the signals at the grid of V16A. This signal is coupled to the Wien bridge circuit through C30 (fig. 7-4). The voltage is developed across both the resistive half of the bridge comprising R71, R72, and R73 and the reactive section consisting of C31, R74, R75, R76, R77, and C32.

With the service selector switch, S4, in the HAND or MACH position, R74 is shorted out through contacts 1 and 7 or 8 of S4D, and R76 is grounded through contacts 6 and 8 or 9 of S4C. Therefore, in hand-key or machine-key operation, the reactive half of the bridge is composed of C31, R75, R76 and C32.

Since R73 is the cathode resistor of V16A, a degenerative feedback voltage, E2, is obtained across R73 in the resistive half of the bridge. A regenerative feedback signal, E1, obtained in the reactive half of the bridge is applied to the grid of V16A. At a frequency determined by the components of the reactive half of the bridge the regenerative voltage, E1, will be in phase with the signal at the grid of V16A. This voltage will be equal to or slightly larger than the degenerative voltage, E2.

The relative amplitude between E1 and E2 is controlled by the setting of the AUDIO OSC

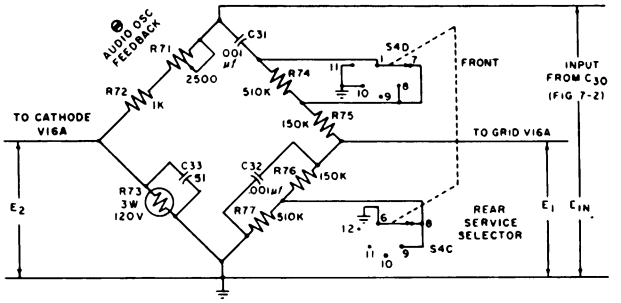

Figure 7-4.—Wien bridge circuit, simplified schematic.

FEEDBACK control R71, which directly determines the amplitude of E2. At any other frequency, E2 will be out of phase and of a smaller amplitude, so that the degenerative voltage, which is independent of frequency, will suppress these frequencies. The resonant frequency of the circuit, when all the components of the reactive half are used, is 200 cycles. This frequency is used during FSK operation. In hand-key and machine-key operation, with R74 and R77 removed by the action of S4 (as discussed), the resonant frequency becomes 1,000 cycles. In facsimile and phone operation the regenerative feedback in the reactive half of the bridge is grounded through contacts 1 and 10 or 11 of S4D, which disables the circuit.

In the cathode circuit of V16A, R73 (a ballast lamp) is used instead of a resistor to stabilize the amplitude of the oscillation. If the amplitude tends to increase, the current through the lamp also increases. The increase in temperature of the lamp filaments causes the lamp resistance to increase, thereby increasing the amount of degenerative feedback. The gain of V16A (fig. 7-2) is reduced in such a manner as to maintain the output voltage at a nearly constant amplitude. If the amplitude of the voltage is kept small, the oscillator tube will operate in the linear portion of its characteristic curve, producing a sinusoidal output.

The output of the oscillator is taken from the plate of V16B and is coupled through C37. In FSK operation the 200-cycle signal is fed to a FSK PHASE MOD control (not shown on the schematic). The control contains an on-off switch that controls the application of the 200-cycle signal. The voltage developed across this potentiometer may be adjusted to produce a one-radian phase shift in the output of the frequency-shift oscillator, varying around the fundamental

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output frequency of 200 cps. The amount of shift introduced into the frequency-shift oscillator is multiplied throughout the remaining units of the RFO. The resulting wobbulating action presented by the 200-cycle shifting causes the transmitter output frequency to shift accordingly.

In hand-key or machine operation, the 1,000-cycle output is coupled through resistor R78 and contacts 6R and 8R or 9R of the service selector switch S4D to the grid of the sidetone amplifier in the audio chain. In hand-key operation a parallel connection, through contacts 8 and 6 of S4F, is made to keying relay, K6 (contact 1R). In space, K6 is not energized, and the 1,000-cycle output is grounded through contacts 2R and 1R of K6. In mark, K6 is energized and the ground is removed, allowing the signal to be applied to the grid of the sidetone amplifier, V4B.

Power Supply in LLRM (+250 V)

The power supply contained in the LLRM provides a regulated +250 volts, which is used in the critical keying and audio circuits of the LLRM and RFO. The power supply consists of rectifier V22, regulator V21, regulator control tube V19, and voltage regulator V20. The a-c power is supplied from the low-voltage power supply through the main power switch. The circuits contained in this power supply are generally the same as those considered in Chapter 3 of *Basic Electronics*, NavPers 10087.

-12-Volt Power Supply

A metallic bridge rectifier, CR2 (fig. 7-2) is used in the -12-volt power supply and is connected so as to yield a negative output. The rectifier receives a 16.5-volt rms signal from terminals 15 and 16 of T4. The output is filtered by L1, C35, and C36, with R79 acting as a bleeder resistor. The -12-volts is supplied to the dynamic-carbon switch, S2, for use as operating voltage for the carbon microphone. The output is also fed to the remote radiophone unit where it serves as the operating voltage for a carrier-on indicator, the control relay, and the carbon microphone.

HIGH-LEVEL RADIO MODULATOR

During phone operation at the 500-watt level, an audio signal of 250 watts is required to amplitude-modulate the plate of the RFA power amplifier. This is accomplished by feeding a 6-watt signal from the LLRM to the HLRM (fig. 7-5) where it is amplified to the required level. During hand-key, machine-key, FSK, and FAX operation, the plate voltage for the RFA power amplifier is routed through the HLRM. In the AN/SRT-16 the output of the HLRM is fed to the RFA in one transmitter group only. This means that only one transmitter group may operate at the 500-watt level. The AN/SRT-15, which has only one transmitter group, may employ the HLRM in 500-watt phone operation. The HLRM is not used with the AN/SRT-14 because this transmitter operates at the 100-watt level only.

PUSH-PULL AMPLIFIERS

During 500-watt operation, the audio modulating signal from the LLRM is fed to the control grids of the push-pull amplifiers, V1 and V2 in the HLRM (fig. 7-5). The signals are equal in amplitude but 180 degrees out of phase, making possible the push-pull amplification. The amplified signals are applied across the output modulation transformer, T1. Plate voltage (+2400 volts during phone) is delivered to the center tap of the transformer primary. The GROUND FOR 500 W PHONE line is supplied from the medium voltage power supply (discussed later) to relay K1. If the -24 V after standby line is activated, K1 energizes. This action supplies the screen grids of V1 and V2 with a +360-volt potential. The potential is applied through contacts 2L and 3L of the now energized K1.

R1 is a dropping resistor for the +350-v screen indicator, I1, which indicates the HLRM screen supply is present. In phone operation the -24 V after standby line is energized by the press-to-talk button such that it controls the operation of K2 and the application of the screen voltage to the HLRM.

In phone operation, +2400 volts from the HVPS is applied to the plates of the power amplifiers in the RFA through T1 in the HLRM. The 250-watt audio signal, which also appears across this transformer, is applied to the PA plates. Protection against excessively high voltage across the secondary of T1 is provided by spark gap E1. In other than phone operation at the 500-watt level, +3000 volts is applied directly to the PA plates through the shorting contacts 4 and 5 of relay K2 in the HLRM. The shorting action takes place in the deenergized condition of the relay. K2 becomes energized through contacts 3R and 4R of energized relay

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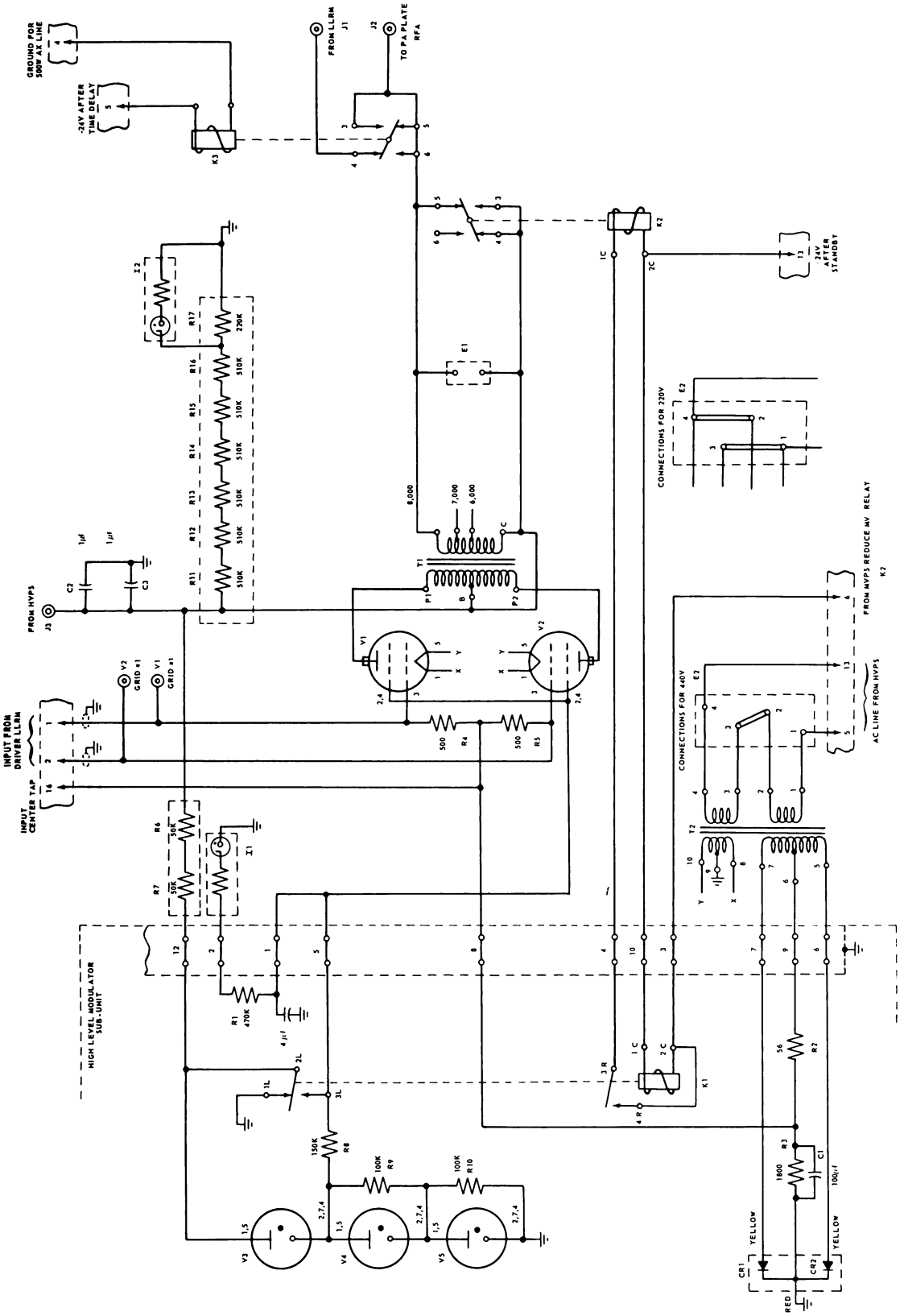

Figure 7-5.—High-level radio modulator, simplified schematic.

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K1. The latter relay energizes when the press-to-talk button is depressed, activating the -24 V after standby line. When K2 energizes and removes the short across the secondary of T1, the transformer is placed in series with the high voltage supplied from the HVPS.

When the transmitter is operated at the 100-watt level, the GROUND FOR 500 W AX line is open as shown in figure 6-3, chapter 6, and K3 is not energized. Therefore, the LLRM connected to J1 passes its output through contacts 4 and 5 of K3 and to J2, which connects to the power amplifier plates in the RFA. In 500-watt operation, the GROUND FOR 500 W AX line is completed, and K3 is energized after the time delay. This action applies the output of the HLRM to the power amplifiers through contacts 6 and 3 of energized relay K3. In the AN/SRT-16, the output of the HLRM to the PA plates (as described) occurs in one transmitter group only. In the other group, the plate potential and the modulating voltage is received from the LLRM directly. As the output of the HLRM is a permanent interconnection, the transmitter group to which this output is fed in the AN/SRT-16 will be the group that can be placed in 500-watt operation.

-50 Volt Bias Supply

In order to generate a bias voltage to operate the audio amplifiers at the desired class AB2 point on its characteristic curve, a -50-volt bias source is included in the HLRM. This supply is obtained from terminals 5 and 7 of T2, using metallic rectifier CR1 and CR2 connected as a negative full-wave rectifier. The rectifier output voltage is taken from terminal 6 of T2 through the single section R-C filter comprised of R2 and C1. The -50 volts is applied to the center tap of the 6-watt secondary of the driver output transformer as explained in the discussion of the LLRM. The -50 volts is also applied to the junction of the audio amplifier grids in the HLRM, through resistors R4 and R5, respectively. These resistors limit the amount of grid current drawn, thereby limiting the amount of distortion.

Transformer T2, through secondary terminals 8 and 10, also supply the filament voltage for the amplifier tubes, V1 and V2. T2 has two primary windings that may be connected in series for 440-volt, single-phase, 60-cycle input, or in parallel for 220-volt input by means of the links on E2.

Regulated +360-Volt D-C Supply

The screen grids of the HLRM receive a regulated +360 volts which is obtained from the voltage divider action of R6, R7, and voltage regulator tubes V3, V4, and V5. The regulator tubes are shorted to ground through contacts 1L and 2L of deenergized K1 during 100-watt operation. During 500-watt operation, relay K1 energizes, removing the ground and applying +360 volts to the screen grid of V1 and V2. R8, R9, and R10 act as a voltage divider in parallel with the regulators to protect the tubes until they fire and take control.

LOW VOLTAGE POWER SUPPLY (LVPS)

The LVPS (fig. 7-6) receives 110-volt, 60-cycle, single-phase power from the ship's supply and delivers this voltage to the MVPS, the RFO, the LLRM, the RFA, and the antenna coupler. The LVPS also supplies the following d-c voltages: +250 volts to the RFO, +300 volts to the RFA and LLRM, -220 volts to the RFA and LLRM, and -24 volts for motor and control circuit functions.

INPUT CIRCUIT OF THE LVPS

The input circuit receives its voltage through receptacle J1 (fig. 7-6). When emergency switch S1 is closed, power is applied through the respective fuses to the start-stop circuit and main power transformer. Power is also applied through fuses and the cabinet heater switch, S3, to space heaters in the cabinet (not shown). The heaters maintain the cabinet at a constant temperature to assure frequency stability.

Application of Power

With the emergency switch closed and the start button on the main power switch, S2, depressed, the master control relay, K4, will be connected across the line. With K4 energized, contacts L3 and T3 close so that when the start button is released, R1 and C1 will be added in series with the relay winding. The addition of these components provide a series impedance with the winding, which allows adequate holding current to keep the relay energized during power on condition. The main power indicator lamp, I2, is also connected across the line through contacts L3 and T3 of K4.

Contacts L1, T1 and L2, T2 of energized relay K1 apply 110-volts a-c to the main power

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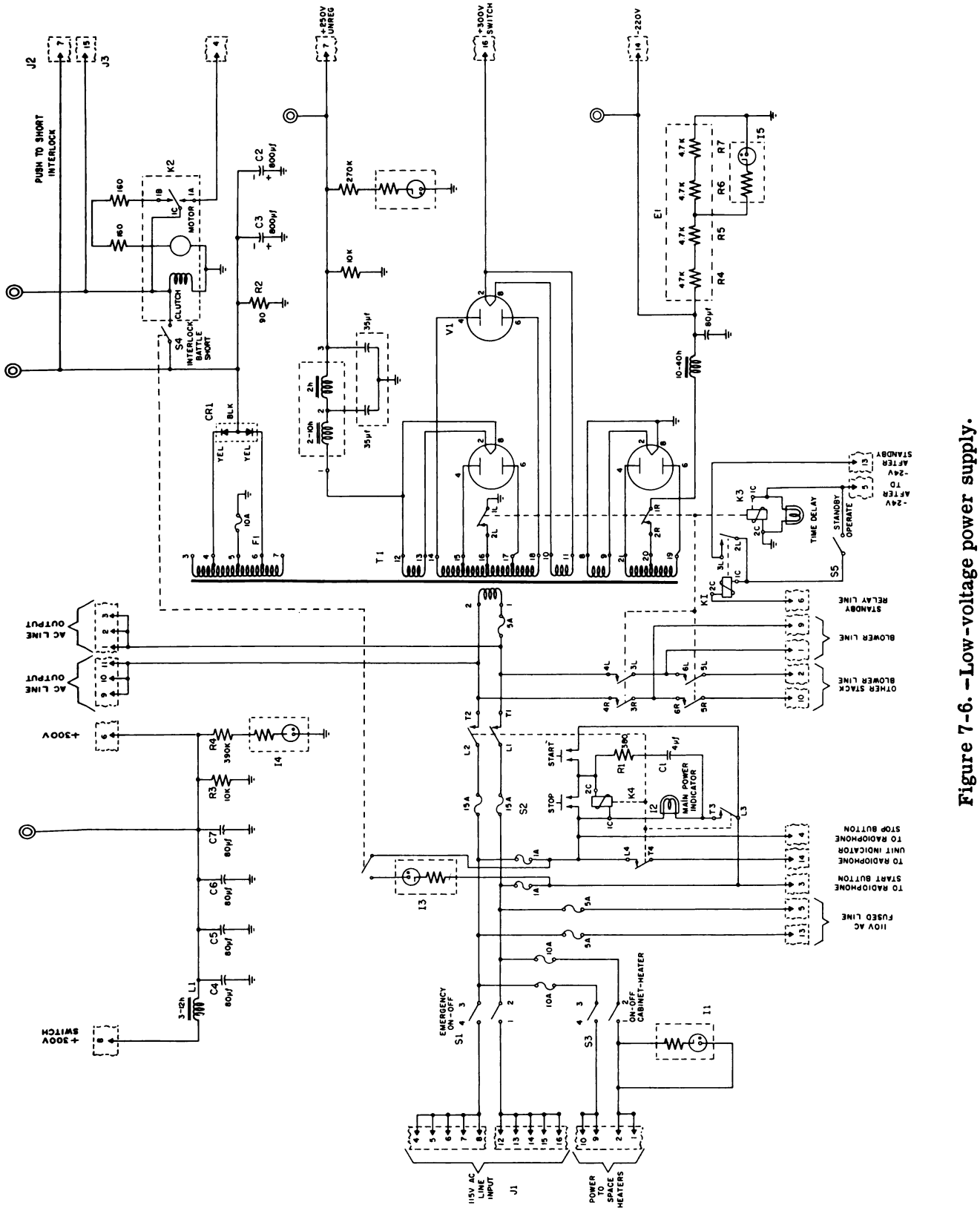

Figure 7-6.—Low-voltage power supply.

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transformer, T1, and to the a-c line output terminal. From this terminal, a-c is applied to the RFO, LLRM, MVPS, RFA, and the antenna coupler. In addition to this action, a-c is applied to contacts 4L and 4R, respectively, of relay K3. If K3 is energized, the a-c potential is advanced to the cabinet blowers (not shown).

When a remote radiophone unit is used, contacts L4 and T4 of energized relay K4 connect a line to the power indicator in this unit.

Removing Power

To turn off the transmitter group, depress the stop button of the main power switch, S2, which places a short across K4. This causes K4 to deenergize. Depressing the stop button on the LVPS does not place a short across the line because of the presence of R1 and C1. With K4 deenergized, power is removed from T1 and all the circuits previously energized by this relay, as discussed above.

-24-Volt D-C Supply

The -24-volt d-c supply is taken from the secondary terminals 4 and 6 (48 volts rms) of T1. The circuit contains a full-wave metallic rectifier, CR1, and the filter comprising C2, C3, and R2. The center tap is grounded through fuse F1.

The output of the filtered -24 volt supply is fed to the time delay relay, K2, through interlocks. The interlocks are contained in the LVPS, MVPS, RFA, LLRM, and RFO. In the AN/SRT-15 and AN/SRT-16 interlocks are also contained in the HVPS and HLRM. If all the interlocks are closed they form a short, which exists between J2 and J3 parallel with the interlock battle short, to energize the time delay relay.

The chassis of the major units of the AN/SRT-14, 15, and 16 may be pulled out on a slide and rail arrangement without disconnecting the cabinet wiring. A drawer interlock is contained in the slide path, which becomes deenergized when the drawer is withdrawn. An interlock battle short, S4, bypasses the drawer interlocks when it is in the ON position by connecting the -24-volt supply directly to the time delay relay. With this switch in the ON position, the interlock battle short indicator, I3, will be energized by 110 volts a-c, through the ganged contacts of S3. The interlock battle short switch should be used in emergency cases only. Before any major chassis is removed from the cabinet, S4 should be placed in the OFF position. Failure to do this will mean that the normal action of the drawer interlock will be bypassed, and the equipment can be energized with the chassis in the OUT position.

The filtered -24 volts from the -24-volt supply of the LVPS is delivered through the interlocks to the motor and clutch of the time delay relay, K2. After approximately 30 seconds, K2 operates its contacts, which supplies the voltage described as -24 volts after time delay.

The plate voltage for the +300 volt supply is obtained from terminals 14 and 18 of T1. This input voltage is approximately 840 volts rms. The center tap (terminal 16) is grounded through contacts 2L and 1L of K3, which is energized when the 30-second time delay times out. This allows for the rectifier filaments to reach operating temperature before V1 is required to conduct. The rectified output is fed to a single choke input filter (upper left of schematic) through contacts 5R and 6R of relay K4 located in the MVPS (fig. 7-7). K4 is energized when the 500-volt supply in the MVPS is put into operation. Because the 300-volt supply is primarily a screen voltage supply and the 500-volt supply is basically a plate voltage source in the LLRM and RFA, the action of K4 ensures that plate voltage is present when screen voltage is applied, thus preventing damage to the screens of the respective tubes they supply. The filter (fig. 7-6) consists of L1, C4, C5, C6, C7, and R3. R3 is a bleeder resistor. R4 acts as a dropping resistor for the +300-volt indicator, I4. The +300-volt output is rated at 200 ma and is fed to the RFA and LLRM.

+250-VOLT UNREGULATED D-C SUPPLY

The +250-volt unregulated d-c supply (also contained in the LVPS) receives its plate input voltage from terminals 15 and 17 of T1. The grounding of the center tap is controlled by relay K3, which is energized after time delay. The circuit operation is generally the same as that discussed in Chapter 3, *Basic Electronics*, NavPers 10087.

-200-VOLT D-C SUPPLY

The -220-volt d-c supply (fig. 7-6) receives its plate voltage from terminals 19 and 21 of T1. This circuit center tap is also controlled by

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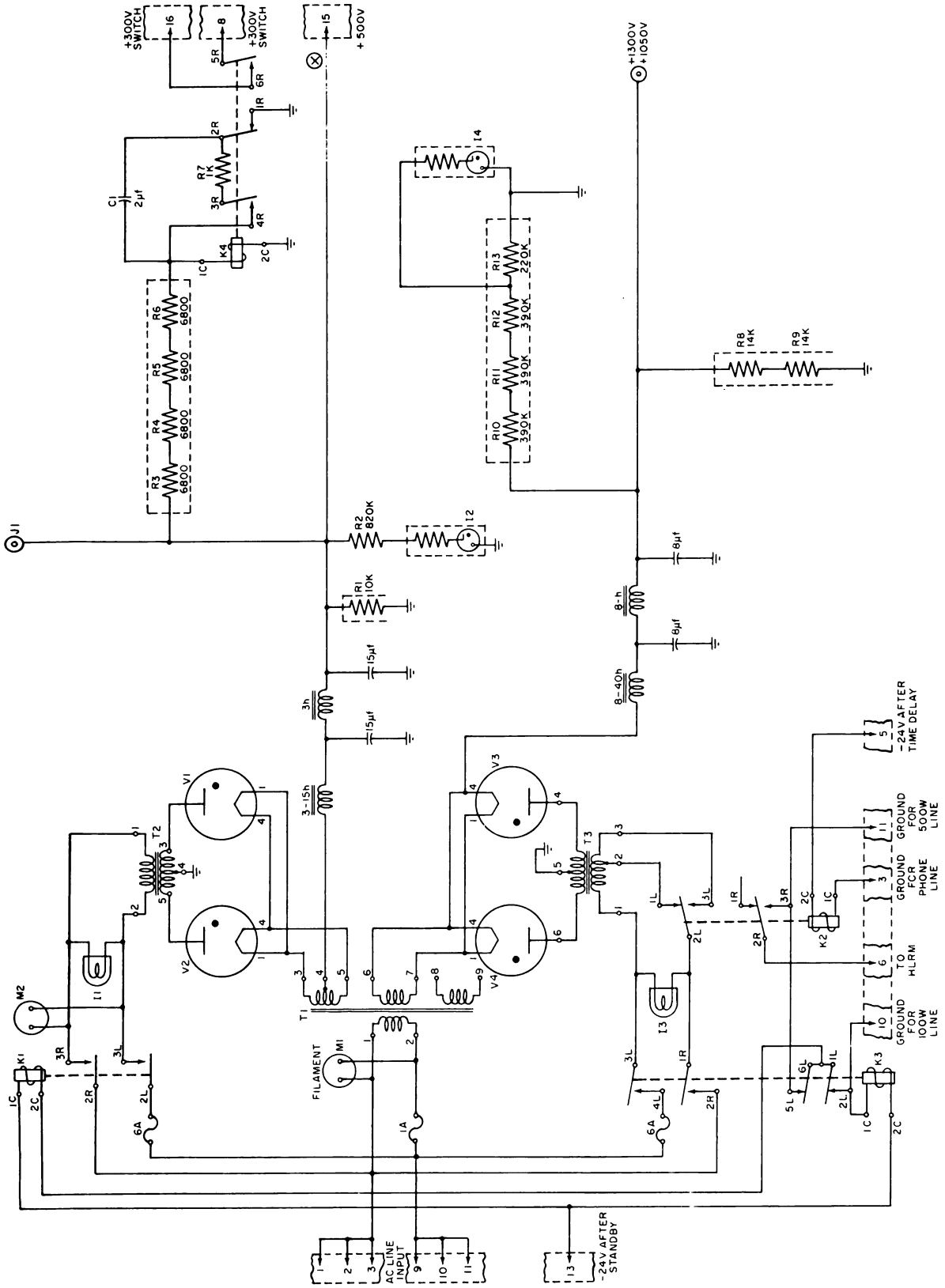

Figure 7-7.—Medium-voltage power supply.

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relay K3, which is energized after time delay. The output voltage is negative because the cathode is grounded and the rectified voltage is taken from the center tap terminal, 20. I5 serves as the -220-volt indicator and, when lighted, denotes the presence of this voltage. E1 is a terminal board used as a mounting for voltage divider resistors R4, R5, R6, and R7.

MEDIUM VOLTAGE POWER SUPPLY (MVPS)

The MVPS (fig. 7-7) receives its input voltage from the LVPS, and delivers +500 volts d-c to the LLRM and RFA and, in addition, +1050 or +1300 volts d-c to the RFA. In addition, the MVPS contains a switching relay to provide assurance that the +500 volts will be available before the +300 volts is applied.

INPUT CIRCUIT OF MVPS

The a-c input to the MVPS is delivered to the filament transformer, T1 (fig. 7-7). One secondary of T1 supplies filament voltage to the +500-volt rectifiers, V1 and V2. The other winding supplies power to the filament of V3 and V4. At the same time a-c power is delivered to the contacts of relay K1, which must be energized before power is applied to the +500-volt plate voltage transformer, T2. V1 is energized from the -24 V after standby line and receives ground through the contacts of relay K3.

If the transmitter is operating at the 100-watt level, and the standby-operate switch, S5, in the LVPS is in the OPERATE position, the -24 V after standby line will be energized after the time delay is complete. Since the GROUND FOR 500 W line is open, the grounding circuit for K3 will be completed through the GROUND FOR 100 W line (fig. 6-3, chapter 6). Therefore, K3 will energize and K1 will energize through contacts 1L and 2L of K3 to the GROUND FOR 100 W line. Energized K1 supplies plate voltage for the +500-volt d-c supply through its contacts, 2R, 3R, and 2L, 3L, and plate transformer T2.

It should be noted that relay K1 can become energized in 500-watt operation through contacts 5L and 6L of K3 and the ground placed on the GROUND FOR 500 W line. In 500-watt operation the 500-volt d-c supply becomes primarily a screen voltage source. Therefore, the transfer of ground is accomplished, providing assurance that the screen voltage will not be supplied before the high voltage plate supply, which is also under control of the GROUND FOR 500 W line.

If the standby-operate switch, S3, in the LVPS is in the STANDBY position on either 100-watt or 500-watt operation, the +500-volt d-c supply will not be energized because the -24 V after standby line (fig. 7-6) will be open and K1 cannot become energized. This, in turn, will prevent the application of a-c power to the plate transformer.

When lighted the indicator lamp, L1, denotes the presence of the a-c plate input, which is approximately 620 volts rms. The rectified output of V1 and V2 of the MVPS is filtered by a double choke input filter. R1 acts as the bleeder resistor, and R2 is a dropping resistor for the 500-volt output indicator, I2. The output is 500 volts and rated at 364 ma, which is supplied to the LLRM and RFA. J1 provides a test point for measuring the +500-volt supply.

Two elapsed time indicators (TOTAL HOURS FIL M1 and TOTAL HOURS PLATE M2) are provided to indicate total time that filament and plate power is on. The filament transformer receives power as soon as the start button in the LVPS is pushed regardless of the position of the standby-operate switch in the LVPS. T2, the plate transformer, receives power only in the operate condition. Therefore, the reading of M2 will indicate the total time that the plate power is on.

+500-VOLT D-C SUPPLY

When the +500-volt d-c supply comes on, its output is applied across resistors R3, R4, R5, and R6 in series with the 300-volt switch relay, K4. The relay is shunted by capacitor C1 through its normally closed contacts, 2R and 1R, which makes relay K4 slow operating. When K4 operates, C1 is removed from the circuit and resistor R7 is placed across C1 through contacts 3R and 4R, providing a discharge path for the capacitor. When K4 is energized, contacts 5R and 6R close, completing the path to the +300-volt filter circuit in the LVPS.

+1050 to +1300-VOLT SUPPLY

The +1050 to +1300-volt supply is used only during 100-watt operation. +1300 volts is supplied to the RFA for all modes of operation except phone when it is reduced to +1050 volts. When relay K3, whose operation has been

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explained, becomes energized, 110-volts a-c is applied to its contacts, 2R, 1R and 4L, 3L. If the transmitter is in other than phone operation, the GROUND FOR PHONE line is open, and relay K2 cannot become energized. Therefore, the a-c power is applied to the plate transformer through contacts 2L and 1L of K2.

The secondary voltage developed between terminals 4 and 6 of the transformer is approximately 3160 volts rms. When rectified and filtered, this produces the +1300 volts, which supplies the RFA. When the transmitter is in phone operation, the GROUND FOR PHONE line is completed through contacts 11F and 1F of S4E in the LLRM (fig. 7-2), and relay K2 of the MVPS energizes from the -24 V after time delay line. When the relay energizes, the turns ratio is reduced so that the secondary voltage becomes 2480 volts rms. The rectified output voltage is therefore reduced to +1050 volts. The smaller voltage is used during phone operation because the power amplifier plate voltage is amplitude modulated by the audio output of either the LLRM or HLRM. If the voltage were not reduced, the peaks of the amplitude modulated plate voltage of the power amplifier would exceed its peak voltage rating.

Since relay K2 energizes only during phone operation, the GROUND FOR 500 WATT line is completed through contacts 2R and 3R of K2 (when the latter is energized). Thus, an additional method of controlling the output of the HLRM during phone operation is obtained.

When K3 energizes, it also supplies 110-volts a-c to the 1300-volt primary indicator, I3. The indicator is not controlled by relay K2, and therefore indicates power applied to the plate transformer, T3, for either the +1050 or +1300-volt output.

The rectified output is taken from the parallel filaments of V3 and V4 and is filtered by a double-choke input filter. R8 and R9 are bleeder resistors. R10, R11, R12, and R13 form a voltage divider to which the 1300-volt indicator, I4, is connected. The output is rated at +1050 volts at 150 ma, and +1300 volts at 180 ma, and is applied to the RFA.

HIGH-VOLTAGE POWER SUPPLY (HVPS)

The HVPS (fig. 7-8) is used only in the AN/SRT-15 and 16 for 500-watt transmission. It receives either 220-volt or 440-volt a-c, 60-cycle, three-phase input. The HVPS supplies a d-c output of +3000 volts or +2400 volts, which is supplied to the HLRM.

INPUT TO HVPS

Power to the HVPS is controlled by the booster emergency switch, S1. The three-phase line supplies voltages to both the plate transformer, T2, and to the filament transformer, T1. The filament transformer is connected across one phase of the input. The fuses in this power supply for 440-volt input are rated at 3.5 amps, and for 220-volt input a rating of 6.25 amps is required. Power is connected to filament transformer T1 through terminal board E1. When the input voltage is 440 volts, a link is placed between terminals 2 and 3 of E1, which places the two windings in series across the line. When 220-volt input is used, the links are placed between terminals 1 and 3 and between 2 and 4, which places the two transformer primaries in parallel. In either case, the voltage across each of the two primaries is 220 volts.

The motor of the time delay relay, K4, is connected across terminals 3 and 4 whose potential is always 220 volts. With S1 closed, the voltage applied to the motor causes the time delay relay to time out in about 30 seconds, which closes its two sets of contacts. Contacts 5 and 7 place the time delay indicator I1 and dropping resistor R1 across the 220 volts.

All of the components of the 500 W disable line must be properly activated before the transmitter can be placed in 500-watt operation. Contacts 2 and 8 of the time delay relay, K4, are situated in the 500 W disable line. Reference to figure 6-3, chapter 6, will show that relay K1 (also in the 500 W disable line) controls the GROUND FOR 500 W line. When K4 times out, the GROUND FOR 500 W line is completed, and if the standby operate switch, S5, in the LVPS is in the operate position, the -24 V after standby line is energized. Thus, the high voltage plate relay K2 is operated. The plate voltage to the HVPS is controlled in this manner so that the filaments of the rectifier tubes, V1 through V6, may reach the proper operating temperature before the high voltage is applied.

When contacts L1, T1; L2, T2; and L3, T3 of K2 close, three-phase power is connected to T2 and to the primary indicator lamps I2, I3, and I4, which are wye connected across the phases and indicate the presence of voltage at the primary of the plate transformer. Resistors

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R2, R3, R4, and R5, R6, R7 are dropping resistors for the indicators. With 220-volt input, R5, R6, and R7 are shorted out by connecting links between terminals 1 and 2, 3 and 4, and 5 and 6 on E2.

HIGH-VOLTAGE POWER SUPPLY OUTPUT

Relay K1 is energized during 500-watt operation, and it reduces the output of the supply from +3000 volts at 355 ma to +2400 volts at 560 ma by decreasing the turns ratio of the plate transformer, T2. The secondary of T2 is delta connected. Each winding develops 2260 volts rms when the transmitter is in c-w operation and 1835 volts during phone transmission. Six half-wave rectifiers, V1 through V6, are connected as a three-phase, full-wave rectifier. When the power is applied, two of the rectifier tubes will be conducting at any given time.

Let us consider the action at one instant in the rectifier operation. Assume that the point HV1 is positive momentarily, and the point HV2 is negative for the same instant. Two rectifiers V6 and V2 are conducting such that the electrons flow from HV2 (negative) through V2, to ground, through the external circuit and filter choke, L1, and back through V6 to HV1. At another instant, the polarity reverses so that HV1 is negative and HV2 positive. In this condition, rectifier V3 conducts, passing a current to ground through the load and back through V5. In this manner, two current peaks are produced for each cycle input. The other two phases also produce two current peaks per cycle, but at different times because of the 120° phase angle between each phase voltage. Six current peaks pass through the load for each input cycle, providing a ripple frequency of 360 cps.

The filter capacitors, bleeder resistor, and 3000-volt indicator, which form an additional portion of the filter circuit with choke L1, are connected from J1 on the HVPS to J3 in the HLRM (fig. 7-5). This has been done because of space limitation in the HVPS. Because of the high ripple frequency, the filter capacitors, C2 and C3, can be smaller than would be required for a single-phase, full-wave filter circuit. R11 through R17 act as a bleeder, and a voltage divider from which the 3000-volt indicator, I2, receives its operating voltage.

There are four filament secondaries of T1 (fig. 7-8), which supply the high-voltage rectifier tubes. Three of these windings are rated at 5 amperes, and the other (connected between terminals 11 and 12) is rated at 15 amperes. The 15-ampere winding supplies rectifiers V4, V5, and V6. Each rectifier (V1, V2, and V3) employs a separate 5-volt filament winding.

The +3000-volt output of the HVPS is supplied to the HLRM in 500-watt operation during c-w, frequency-shift, and facsimile operation. During phone transmission at the 500-watt level, the +2400-volt output is supplied to the HLRM.

RADIOPHONE UNIT

The radiophone unit is used in conjunction with, but is not a part of the AN/SRT-14, 15, and 16. The unit is usually located at a remote position and contains a start-stop switch for turning the transmitter on or off, jacks for connecting a handset, chestset (or hand key) a volume control for the earphones, and visual indicators for power and carrier.

A complete description of a representative Navy radiophone unit may be found in Chapter 4, Control System, *Electronics Technician 2,* Vol. 2.

QUIZ

1. (a) The electronic keyer circuit (fig. 7-1) of the low-level radio modulator provides the keying voltage to control the RFA and what units of the RFO during amplitude modulation? (b) What unit alone does the electronic keyer circuit control during frequency-shift keying?
2. The audio input signal to the LLRM (fig. 7-2) is received locally through which jacks?
3. By applying the audio signal to the suppressor grid of V1 (fig. 7-2) and the AGC voltage to the control grid, what feature of the AGC is improved?
4. Limiting of the audio signal amplitude to prevent overmodulation of the carrier is accomplished by which tube in the LLRM (fig. 7-2)?
5. The squelch circuit (fig. 7-2) consists of what stages?

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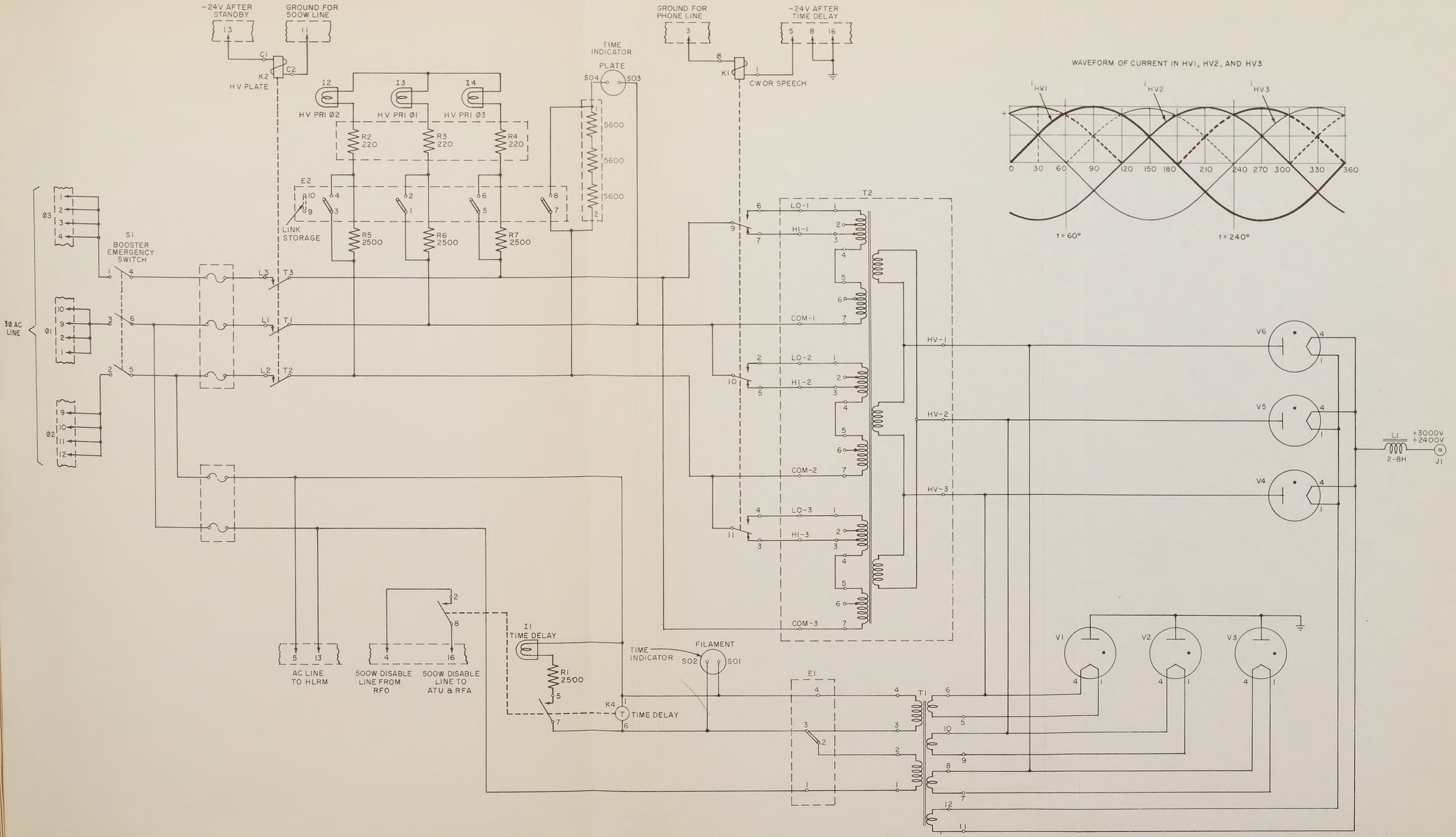

Figure 7-8. –High-voltage power supply.

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6. The output of the sidetone amplifier, V4B (fig. 7-2) is coupled by an output transformer, T2, to what three monitoring circuits?
7. The amount of degeneration in the cathode circuit of the phase inverter, V5B (fig. 7-2) is approximately equal to the amplitude of the input signal for what purpose?
8. V8A and V8B (push-pull cathode followers) act as drivers for what stages of the LLRM (fig. 7-2)?
9. The no-signal grid bias of the power amplifier applied from the cathode followers is approximately what value (fig. 7-2)?
10. In hand-key and phone operation, the keying signals are introduced to the LLRM through the action of what control (fig. 7-2)?
11. What is the purpose of the keying control tube, V12B (fig. 7-2)?
12. The dumping cathode follower provides a low impedance to what stages of the LLRM (fig. 7-2)?
13. Facsimile signals are developed across resistor R70 (fig. 7-2) and fed through what section of the service selector switch to the output terminals?
14. During phone operation at the 500-watt level, an audio signal of what value is required to amplitude modulate the plates of the RFA power amplifiers?
15. In phone operation, +2400 volts from the HVPS is applied to the plates of the power amplifiers in the RFA through what component in the HLRM (fig. 7-5)?
16. In other than phone operation at the 500-watt level, what value of voltage is applied directly to the PA plates through the shorted contacts 4 and 5 of relay K2 in the HLRM (fig. 7-5)?
17. The low-voltage power supply (fig. 7-6) receives 110-volts, 60-cycle, single-phase power from the ship's supply and delivers this voltage to what other units?
18. Depressing the stop button on the LVPS (fig. 7-6) does not place a short across the line because of the presence of what components?
19. The filtered -24 volts from the 24-volt supply of LVPS (fig. 7-6) is delivered through the interlocks to what component?
20. The rectified output of V1 and V2 of the MVPS (fig. 7-7) is filtered by what type filter?
21. The +1050- to +1300-volt supply is used only during what operating level (fig. 7-7)?
22. When the transmitter is in phone operation, the GROUND FOR PHONE line is completed through contacts 11F and 1F of S4E in the LLRM, and relay K3 (fig. 7-7) becomes energized from what source?
23. The high-voltage power supply (fig. 7-8) is used in what two transmitting sets?
24. Power to the HVPS (fig. 7-8) is controlled by what switch?
25. Contacts 2 and 8 of the time delay relay, K4 (fig. 7-8) are situated in what circuit?
26. The +3000-volt output of the HVPS is supplied to the HLRM in 500-watt operation during what modes of transmission?
27. During phone operation at the 500-watt level, what output is supplied to the HLRM?

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CHAPTER 8

PRINCIPLES OF SINGLE-SIDEBAND COMMUNICATIONS

INTRODUCTION

A brief discussion of single sideband is included in Chapter 1 of *Electronics Technician 3*, NavPers 10188-A. However, the subject is treated in greater detail in this chapter. Because single sideband is becoming increasingly important in Navy applications, it is necessary that Navy technicians acquire an understanding of the basic principles of operation. If the technician understands the principles of operation, he will experience little difficulty in utilizing the instruction books or technical manuals in corrective maintenance work.

It is, of course, not possible to give complete coverage in one chapter of reasonable length. Some references that will be of value to the ET are included in *Electronics Information Bulletin*, No. 435 of 18 June 1956. Also, helpful articles appear from time to time in the current literature (for example, in the *Bureau of Ships Journal*, beginning March 1958). The alert technician will make every effort to keep abreast with the advances in the field.

It is the purpose of this chapter to list, first of all, the advantages of single sideband (SSB) and to point out the differences between SSB and the conventional system of amplitude-modulation communication, utilizing both sidebands and the carrier. Next, some of the more important problems in SSB communication are discussed.

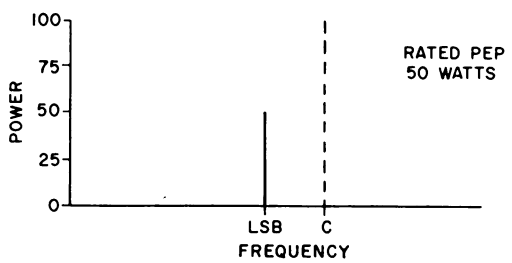
In order to approach the subject of equipment operation in the simplest and most logical manner, a functional block diagram of a single sideband transceiver that has been widely used aboard ship is given first. This is accompanied by a discussion of the functions performed in each of the blocks.

A simplified circuit diagram of the equipment is included next. The various circuits are analyzed to better acquaint the technician with circuit operation. The actual circuit diagram, including the various switching functions, is included in the instruction book that comes with the equipment.

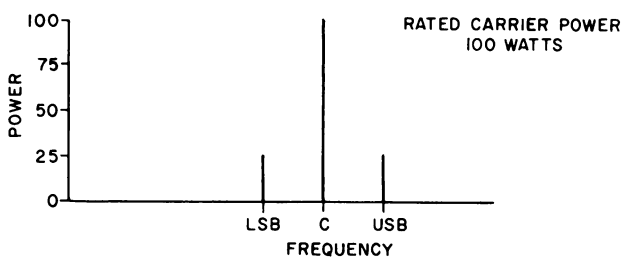
The chapter concludes with a discussion of the various methods of SSB generation, the commonly used types of filters and linear amplifiers, and a brief introduction to shore-based SSB equipment.

ADVANTAGES OF SINGLE-SIDEBAND COMMUNICATIONS

A comparison of the frequency and power relationships between single-sideband transmission and conventional a-m transmission is illustrated in figure 8-1. The SSB system is illustrated in part A of this figure. All of the radiated power is useful in conveying intelligence, and none of the power is in the carrier or the upper sideband. (As was pointed out in



A SINGLE SIDEBAND SYSTEM



B CONVENTIONAL A-M SYSTEM (100% MOD.)

Figure 8-1. –Comparison of Amplitude modulation and SSB.

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NavPers 10188A, only one sideband is needed to transmit intelligence.) Although the lower sideband is transmitted (in the illustration), the upper sideband could have been transmitted just as easily. In some systems (AN/WRT-2, AN/WRR-2 and shore-based equipment), both sidebands may be utilized independently.

In the familiar a-m system of communication (fig. 8-1, B), the radiated signal includes the carrier and an upper and a lower sideband frequency for each frequency in the modulating signal. For example, if a 1-mc carrier is modulated by a 1-kc tone, the radiated signal will include the 1-mc carrier, the lower sideband frequency (1 mc - 1 kc = 999 kc), and the upper sideband frequency (1 mc + 1 kc = 1001 kc). If the modulating signal contains many frequencies, there will be, of course, many frequencies in the sidebands. In this system of transmission, none of the transmitted intelligence is contained in the carrier, and therefore the power put into the carrier is wasted insofar as transmitting intelligence is concerned. Likewise, because duplicate information is contained in each of the two sidebands, the intelligence content of the transmitted signal could be recovered from one sideband only.

In a conventional a-m system where both sidebands and the carrier are transmitted, the power in the sidebands is dependent upon the amount of modulation. For 100% modulation the power in the sidebands is equal to one-half that in the carrier. Thus, a conventional a-m transmitter with 100-watts carrier power will have 50 watts in the sidebands (25 watts in the upper sideband and 25 watts in the lower sideband) at 100% modulation, making the total power transmitted 150 watts (fig. 8-1). It can be seen, then, that two-thirds of the total radiated power in a conventional a-m system (assuming 100% modulation) is in the carrier and is therefore not useful in conveying intelligence.

When the r-f signal is demodulated in the conventional a-m system, the audio output is a combination of the upper and lower sidebands. In this conventional type of detection the audio output is proportional to the power contained in the two sidebands.

In a single-sideband system, only one sideband is transmitted and therefore the audio output of the SSB receiver is proportional to the power contained in the one sideband.

It therefore becomes apparent that an SSB transmitter and an a-m transmitter will perform equally (same signal to noise ratio) under ideal propagating conditions, if the total sideband power of the two transmitters is equal. Considering the relationship between sideband power and carrier power in a conventional a-m system, it is evident that an SSB transmitter will perform as well as an a-m transmitter of twice the power rating under ideal propagating conditions. Thus, a single-sideband transmitter rated at 50 watts will produce the same signal intelligence level at a receiver as a conventional a-m transmitter rated at 100 watts (see figure 8-1).

As propagating conditions become less than ideal the SSB system will show even a greater advantage over an a-m system. An a-m transmission is subject to deterioration under poor propagation conditions, because all three components of the transmitted signal (the upper sideband, the lower sideband, and the carrier) must be received exactly as transmitted to realize perfect reception. Because there is only one component in the transmitted signal for an SSB system, it is not so affected by poor propagating conditions. Studies have shown that the SSB system will give from zero to nine decibel improvement under various conditions of propagation when the total sideband power in SSB is equal to that in amplitude modulation (fig. 8-2).

Note that under the average conditions (good), the SSB system shows about a 3 db (4 fold) advantage over the a-m system. In other words, in normal use, an SSB transmitter rated at 100

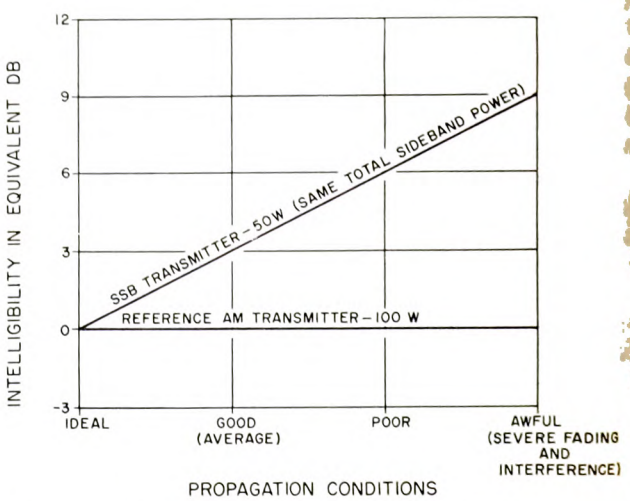

Figure 8-2. –Comparison of Amplitude modulation and SSB under varying propogation conditions.

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atts (PEP will give equal performance with
n a-m transmitter rated at 400-watts carrier
ower.

As far as bandwidth is concerned (assuming
ne sideband only), the SSB system requires
nly about one-half the frequency spectrum that
ie conventional a-m system requires.

The advantages of SSB over the conventional
-m system may be summarized as follows:

1. The SSB transmitter will perform as well
s an a-m transmitter of twice the power rating
nder ideal propagating conditions. Under aver-
ge conditions there is also an additional 3 db
dvantage of a SSB system over an a-m system
aving the same sideband power.

2. If only one sideband is used, the SSB sys-
em requires about one-half as much r-f spec-
rum as the a-m system.

3. The SSB transmitting system uses small-
r units than comparable a-m units because less
ower is required.

4. By virtue of less power in the antenna,
ower voltages are obtained, with attendant re-
luction of potential breakdown.

5. The SSB system is subject to less noise
nterference because the bandpass is narrower.

PROBLEMS INVOLVED IN SINGLE-SIDEBAND COMMUNICATIONS

The advantages of SSB cannot be realized
vithout the use of specially designed components
nd circuitry. First of all, there is the problem
f frequency stability, especially when the car-
ier is totally suppressed. This means that the
scillators in the transmitter and in the re-
eiver must not drift more than a few cycles.
ctually, the permissible frequency variation
or SSB systems is $\frac{1}{100}$ th of that for an a-m
ystem.

In one type of double-sideband generation,
ilters of extreme selectivity are needed. Linear
ower amplifiers, which are difficult to design,
re also needed.

Another problem, when SSB equipment is used
n high-speed aircraft, is that of Doppler shift.
his is especially noticeable at the higher radi-
ted frequencies.

FUCTIONAL BLOCK DIAGRAM OF A SHIPBOARD SSB SYSTEM

A block diagram of an SSB transceiver is
hown in figure 8-3. This is a single-sideband,
uppressed-carrier, transmitter-receiver combination (some examples of SSB systems are the AN/URC-32, the AN/WRT-2, and the AN/WRR-2). The operating frequency of this system is selected from one of four pretuned channels, and the signal is transmitted on the lower sideband only.

The SSB system illustrated in figure 8-3 consists of a transmitter-receiver chassis and a power supply chassis.

As may be seen in the diagram, three crystal oscillators are used on transmission and reception. The three oscillators, and the balanced modulators that they feed, are needed to heterodyne the frequency up to the desired radiation frequency and down to the desired demodulation frequency.

TRANSMITTER SECTION

On transmission, the modulating signal (1 kc is assumed) from the tone oscillator (block 1) or the microphone is amplified in the microphone amplifier (block 3).

From block 3 the signal is fed to block 4, a balanced modulator. Here the 1-kc signal is combined with the 250-kc signal from the crystal oscillator (block 5) to produce lower and upper side-band frequencies (250 kc - 1 kc = 249 kc, and 250 kc + 1 kc = 251 kc). The 250-kc oscillator frequency is suppressed in the balanced modulator. A brief description of the balanced modulator is given in Chapter 1 of *ET 3*, NavPers 10188-A; it is also described later in this chapter.

From block 4 the two sideband frequencies are fed to a mechanical filter (block 6). The mechanical filter is a mechanically resonant device that receives electrical energy, converts it into mechanical vibration, then converts the mechanical energy back into electrical energy (using the magnetostriction principle) at the output. This filter resonates with, and passes, the upper sideband (251 kc), but suppresses the lower side-band (249 kc).

From block 6 the 251-kc signal is fed to the second balanced modulator (block 7). Here the 251-kc signal is combined with the 1150-kc signal from block 8. Two sidebands (899 kc and 1410 kc) are produced, and the crystal frequency (1150 kc) is suppressed. Included in this block is a tuned r-f transformer, which passes the upper sideband (1401 kc) and suppresses the lower sideband (899 kc).

The 1401-kc upper sideband frequency from block 7 is fed to the third balanced modulator

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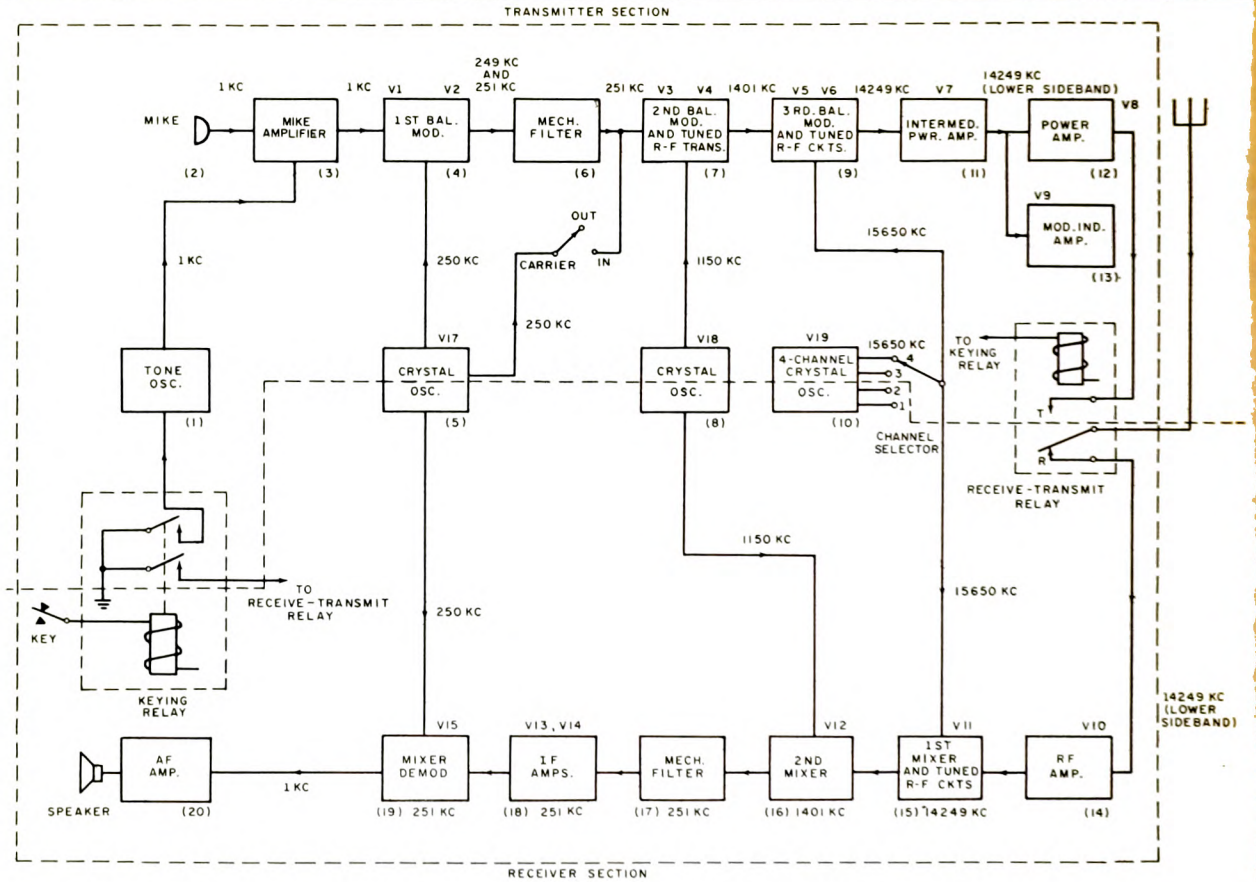

Figure 8-3. –An SSB transceiver.

(block 9) where it is combined with the 15,650-kc signal (assume channel 4 operation) from block 10. The 15,650-kc frequency is suppressed, and the two sidebands (14,249 kc and 17,051 kc) are fed to the tuned r-f circuits in block 9. Only the lower sideband (14,249 kc) is fed to the intermediate power amplifier (block 11) and the power amplifier (block 12).

The lower sideband frequency of 14,249 kc is applied to the antenna when the receive-transmit switch is in the TRANSMIT position. The receiver-transmit relay contacts make electrical continuity between the antenna and receiver at all times except when actually transmitting.

The modulation indicator amplifier (block 13) amplifies the signal from the power amplifier grids and applies it to the peak modulation indicator (a lamp), which indicates modulation peaks.

RECEIVER SECTION

Reception is essentially the reverse of transmission. For simplicity, it is assumed that the incoming signal is 14,249 kc and that the intelligence signal is a 1-kc tone; the nominal carrier is therefore 14,250 kc. Note that with this SSB system the carrier is not transmitted. The 14,249-kc signal is fed (when the receive-transmit switch is in the RECEIVE position) to the r-f amplifier (block 14).

From the r-f amplifier the 14,249-kc signal is fed to the first mixer (block 15) where it is mixed with the 15,650-kc signal from block 10. The output contains the sum and difference frequencies, 29,899 kc and 1401 kc, respectively. The difference frequency (1401 kc) passes through the tuned r-f circuits to the second mixer (block 16).

In the second mixer the 1401-kc signal is mixed with the 1150-kc frequency from block 8 to produce sum and difference frequencies

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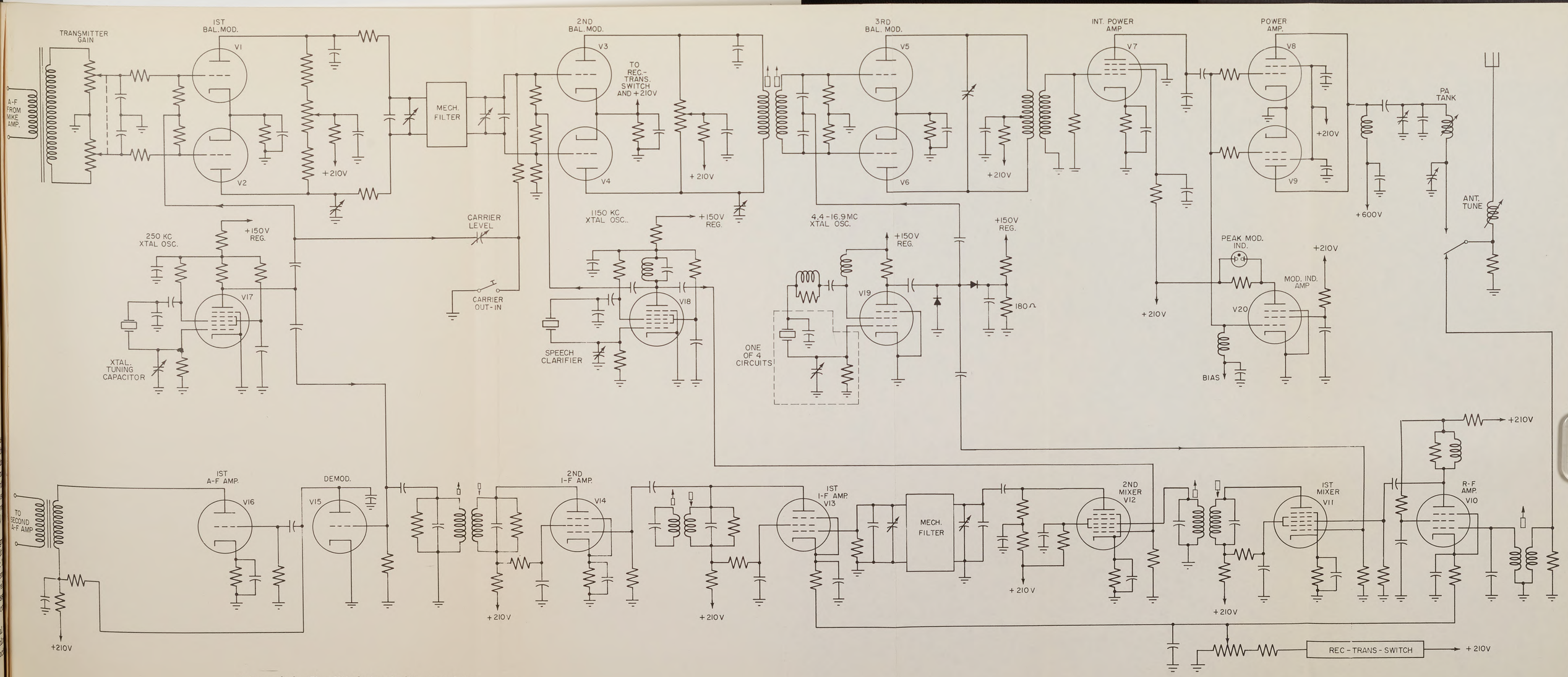

Figure 8-4. –Simplified circuit diagram of an SSB-1 transmitter-receiver chassis.

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2551 kc and 251 kc, respectively. The mechanical filter (block 17) is tuned to 251 kc, and this is the only frequency that is passed through the i-f amplifier to the mixer demodulator (block 19).

In the mixer-demodulator the 251-kc, i-f signal is mixed with the 250-kc frequency from block 5 to produce sum and difference frequencies, 501 kc and 1 kc, respectively. The 501-kc frequency is eliminated, and the 1-kc signal is then amplified in block 20 and applied to the speaker.

TRANSMISSION WITH CARRIER

To make this equipment compatible with a-m equipment now in use, provision is made for inserting a carrier signal after the mechanical filter (at the input of the second balanced modulator, see the transmitter section of figure 8-3).

The 250- and the 251-kc signals are heterodyned in the second and third balanced modulators. The frequency difference between carrier and sideband is maintained in the output the same as in the input to the second balanced modulator. Assume channel 4 operation (15,650 kc). The carrier will then be 14, 250 kc, and the lower sideband (produced by the 1-kc tone) will be 14, 249 kc. The combined signal can be detected by conventional a-m receivers that are capable of being tuned to this frequency.

SIMPLIFIED CIRCUIT DIAGRAM

Figure 8-4 is a simplified schematic diagram of the circuits included in the transmitter-receiver chassis. The audio input section (including the microphone amplifier and the tone oscillator) is treated later; the audio output section is also treated later. Power supplies are treated in *Basic Electronics*, NavPers 10087; and, because the power supply is conventional, it is not covered in detail in this chapter.

CRYSTAL OSCILLATORS

The function of the various circuits in the SSB system have been discussed in terms of the functional block diagram. It is convenient to begin the brief analysis of the circuits in the transmitter-receiver chassis (fig. 8-4) with the crystal oscillators.

Two of the crystal oscillators, V17 and V18, are of fixed frequency (250 kc and 1150 kc), and the necessary crystals are supplied with the equipment. The four-channel-frequency crystals used with V19 are supplied separately. The nominal carrier frequency (of course, the carrier is not transmitted) determines the frequency of the channel-frequency crystals. When channel-frequency crystals are to be selected it is a simple matter to determine the correct crystal frequencies: simply add 1400 kc to the desired nominal carrier frequency for each channel. For example, assume that the channel selector switch is set to channel 4 and that a nominal carrier frequency of 14,250 kc is to be used. The crystal frequency must then be 14,250 kc plus 1400 kc, or 15,650 kc.

The 250-kc crystal oscillator, V17, feeds the first balanced modulator, V1-V2, on transmission and the demodulator, V15, on reception. The 250-kc crystal (as well as the others) is enclosed in an oven where a constant temperature is maintained. As indicated in figure 8-4, the plate of V17 is supplied from the 150-v regulated supply. The crystal is connected (through a coupling capacitor) between grid 3 and grid 1, making, in effect, an electron-coupled oscillator.

The crystal tuning capacitor is used to vary the oscillator frequency over a small range.

Various types of oscillators are treated in Chapter 7 of *Basic Electronics*, NavPers 10087. The crystal may be connected between various tube elements—for example between the first and third grids of a pentagrid tube, between the screen grid and control grid of a pentode, or between the plate and grid or grid and cathode of a triode.

The 1150-kc oscillator, V18, operates in a manner similar to the 250-kc oscillator. However, the tuning capacitor in this instance has a front-panel control, called the speech clarifier. By adjusting this capacitor the operator may bring the oscillator frequency exactly to that of the station he is working. This oscillator feeds the second balanced modulator, V3 and V4, on transmission and the second mixer, V12, on reception.

The channel-frequency oscillator, V19 (a simplified diagram of which is shown in figure 8-5) employs a power pentode and any one of the four crystals. The position of the switch determines which of the crystals will be used. This oscillator feeds the third balanced modulator, V5 and V6, on transmission and the first mixer, V11, on reception.

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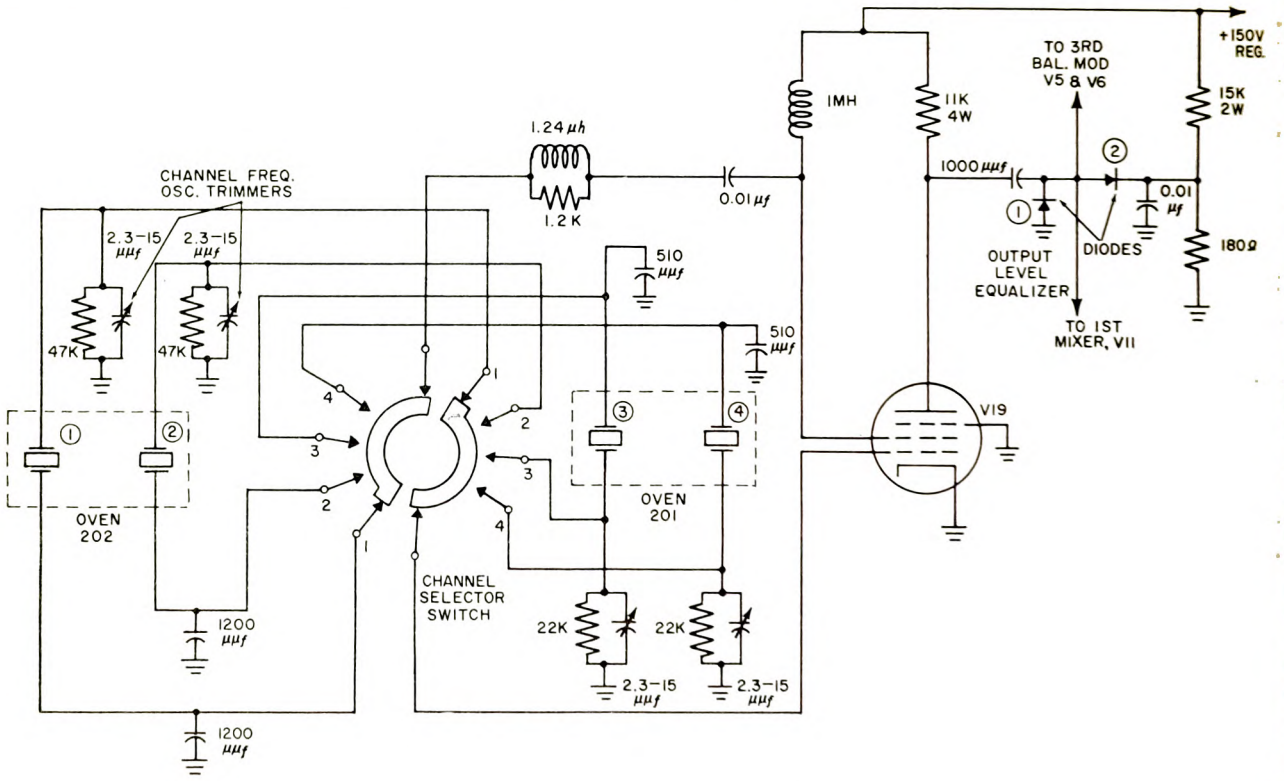

Figure 8-5. –Simplified diagram of the channel-frequency oscillator.

As may be seen in the figure, the crystal is connected between the screen grid and the control grid. Feedback occurs through the 0.01 μf capacitor (coupled to the screen grid) and the parallel combination of the 1.24-μh choke and the 1200-ohm resistor. Grid voltage is developed across a 47,000-ohm resistor for channels 1 and 2 and across a 22,000-ohm resistor for channels 3 and 4. For each position of the channel selector switch a 2.3- to 15-$\mu\mu$f trimmer capacitor between grid and ground is available for adjusting the oscillator circuit to the desired frequency.

The crystal-oscillator, output-level equalizer employed in the output of the channel-frequency oscillator limits the output (fed to the third balanced modulator, V5 and V6 and the first mixer, V11) to approximately 1.8 v, which is the voltage drop across the 180-ohm resistor. The limiting action may be explained by the use of the simplified diagram shown in figure 8-6. In this diagram the 180-ohm resistor has been replaced by a 2-volt battery (2 volts are used to avoid the decimal), and only the plate and cathode connections of the oscillator tube, V19, are shown. The easy direction of electron flow in the diodes is opposite to that of the arrow in the diode symbol.

The input voltage, E_{a-g}, swings 50 v above and 50 v below an assumed value of 100 v. The voltage E_{a-b}, across capacitor C swings 49 v above and 49 v below 99 v because of the action of the 2-volt battery. The voltage, E_{b-g}, fed to the coupling capacitor is a series of pulses that swings between 0 and 2 v. Because of the action of the coupling capacitor and because the diodes are not perfect one-way conductors, the voltage fed to the load is essentially of a sine waveform having a peak-to-peak value of two v.

At 0° it is assumed that E_{a-g} is at its peak value of 150 v; E_{a-b} is at its peak value of 148 v, and E_{b-g} is at its maximum value of 2 v. These values have been established during the previous half cycle.

When E_{a-g} decreases to 149 v, E_{a-b} remains at 148 v, and E_{b-g} decreases to 1 v. When E_{a-g} decreases to 148 v, E_{a-b} still remains at 148 v, and E_{b-g} decreases to 0 v. When E_{a-g} decreases to 147 v, C begins to discharge through diode 2,

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and the discharge continues to the 180° point, thus maintaining point B at ground potential during this portion of the input cycle.

At 180°, E_{a-g} is 50 v and E_{a-b} is also 50 v. When E_{a-g} increases to 51 v, E_{a-b} remains at 50 v, and E_{b-g} increases to 1 v. When E_{a-g} increases to 52 v, E_{a-b} still remains at 50 v and E_{b-g} increases to 2 v. When E_{a-g} increases above 52 v, diode 1 conducts, thus maintaining E_{b-g} at 2 v until the 360° point is reached.

Immediately following 180° the rising voltage, E_{a-g}, is imparted to the load because neither diode is conducting. The rise in voltage across the load is terminated at the instant diode 1 conducts and is limited to 2 v during this half cycle. Similarly, following 360° the falling E_{a-g} voltage change is imparted to the load because neither diode is conducting. At the instant E_{b-g} becomes zero, diode 2 conducts, and the output remains at zero potential during this half cycle.

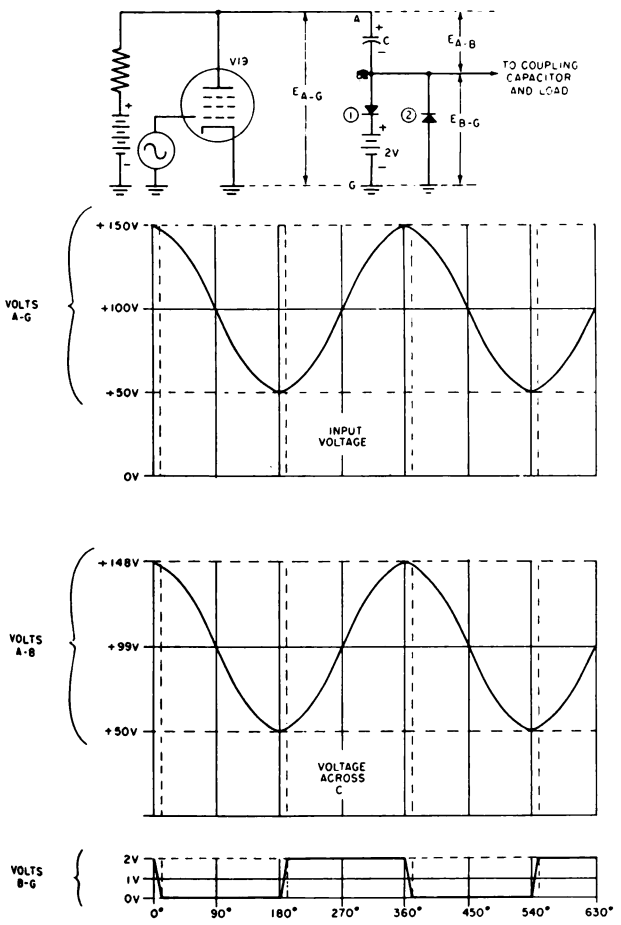

Figure 8-6.—Oscillator output level equalizer.

METHODS OF GENERATING SSB SIGNALS

There are two common methods of generating single sideband signals. One method involves the use of the balanced modulator and a suitable filter. It is commonly called the filter method. The other involves the use of phasing circuits, and is commonly called the phasing method. Both methods are discussed in the following paragraphs.

FILTER METHOD

A simplified circuit diagram of the first balanced modulator (V1 and V2 in block 4, figure 8-3) is given in figure 8-7. The 250-kc signal generated by the crystal oscillator is applied in the same phase to the grids of V1 and V2, and is therefore suppressed in the output. The reason the carrier is suppressed may be explained as follows: (1) at a given instant the crystal oscillator drives both grids an equal amount in a positive direction, (2) plate current then rises an equal amount in both tubes, (3) current flows from both plates toward the movable contact on the balance control, (4) the voltages across both plate loads are equal and opposite, and (5) no difference in potential at the carrier frequency exists across the input to the mechanical filter, and hence the carrier is suppressed.

The amount of carrier suppression depends on the degree of balance that is maintained. Under conditions of perfect balance, the carrier is completely suppressed.

The 1-kc signal fed through the a-f transformer arrives at the grids of tubes V1 and V2 180° out of phase; and, as in any push-pull circuit (see Chapter 6 of *Basic Electronics*, NavPers 10087), a signal appears at the output. However, the audio component is not passed by the filter or any of the other r-f circuits.

The two sideband components (sum and difference frequencies of 249 kc and 251 kc) generated in the circuits of V1 and V2 are not balanced out in the plate circuits and therefore appear in the output. They are not balanced out for the same reason that the audio component is not balanced out. That is, each of these frequencies applied to the grid of V1 is 180° out of phase with the corresponding frequency applied to the grid of V2.

Because only the upper sideband is to be passed to the second balanced modulator (block

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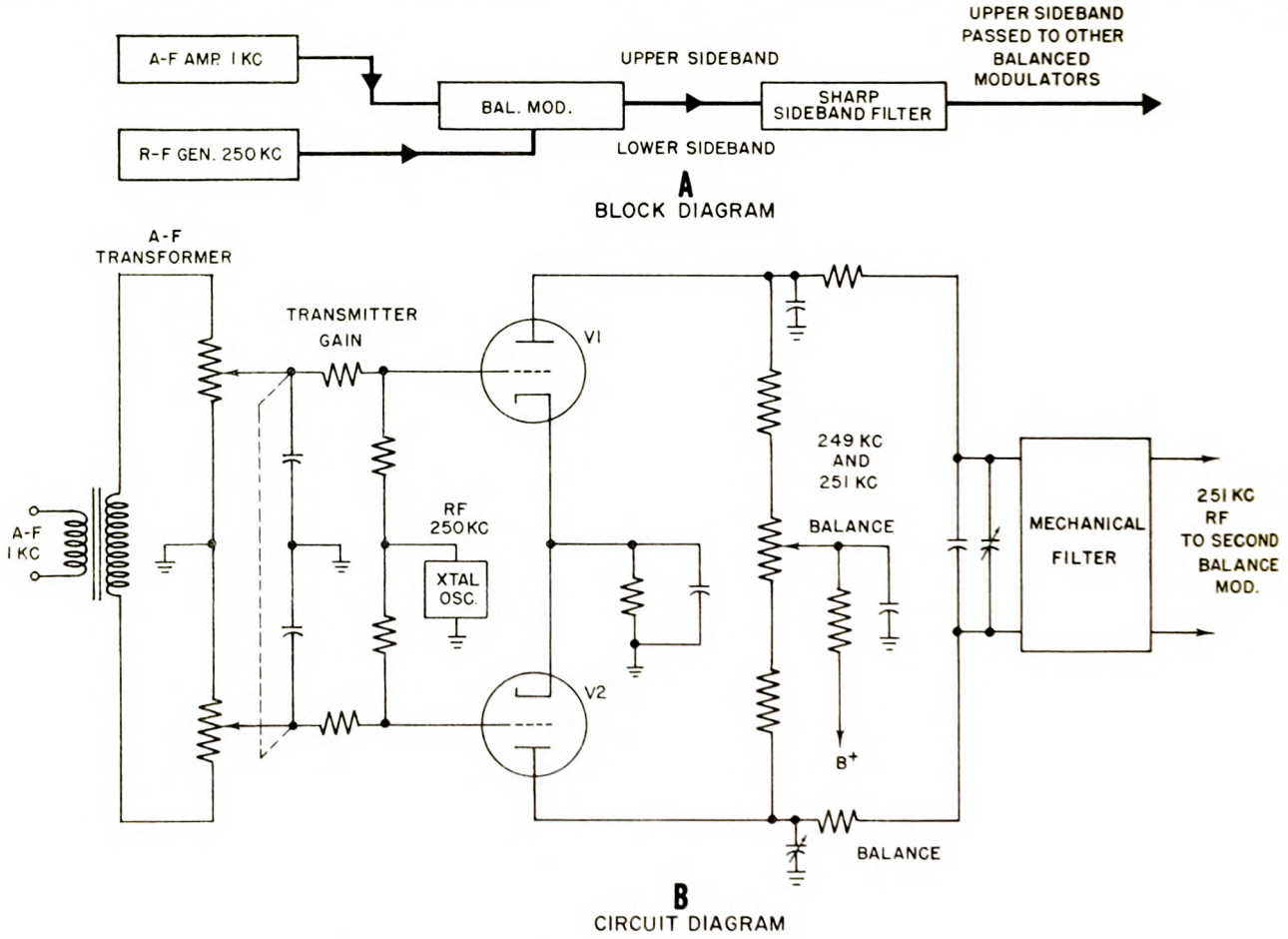

Figure 8-7. –Block and circuit diagram of a balanced modulator.

7, figure 8-3), the mechanical filter is designed to pass 251 kc and to eliminate 249 kc.

PHASE-SHIFT METHOD

The phase-shift method of generating SSB differs considerably from the previously described filter method. In the phase-shift method, two balanced modulators are connected in an equivalent push-pull circuit (fig. 8-8). The input signals are shifted 90° with the result that either the sum or the difference frequencies are cancelled in the output, thereby eliminating the need for narrow band filters and additional stages in cascade. In figure 8-8, A, the upper sideband is suppressed and the lower sideband is radiated; and in figure 8-8, B, the lower sideband is suppressed and the upper sideband is radiated. Actually, the arrangement in part A can be changed to that in part B by means of a switching circuit, and this is the usual method of operation. Thus, the operator has a choice of operating on either the lower or the upper sideband.

Assume that in figure 8-8, A, the modulating frequency is 1 kc and that the r-f carrier oscillator is 5 mc. The a-f signal is fed directly to balanced modulator 1; it is also fed to the a-f phase shifter. From the a-f phase shifter the signal is shifted 90° and fed to balanced modulator 2.

The r-f carrier is fed to balanced modulator 1 with no shift in phase; it is also fed to balanced modulator 2 through the r-f phase shifter.

In balanced modulator 1 the a-f and r-f signals are combined to produce upper and lower sideband frequencies. The upper sideband from balanced modulator 1 is 180° out of phase with

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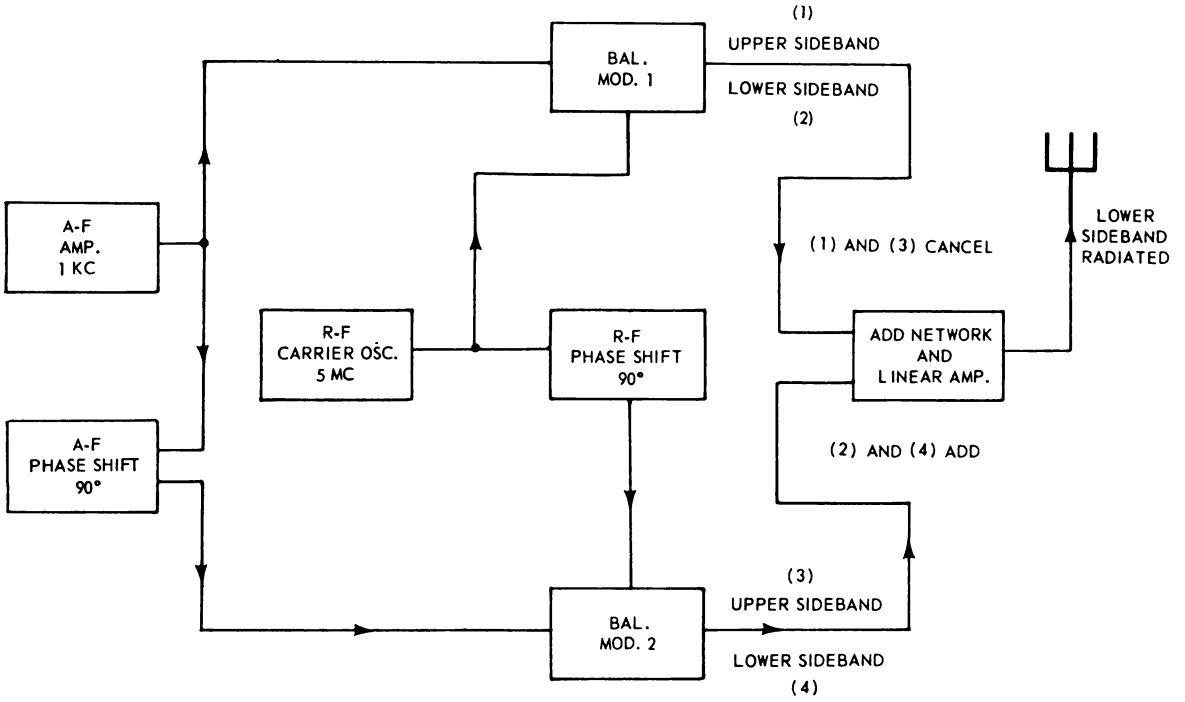
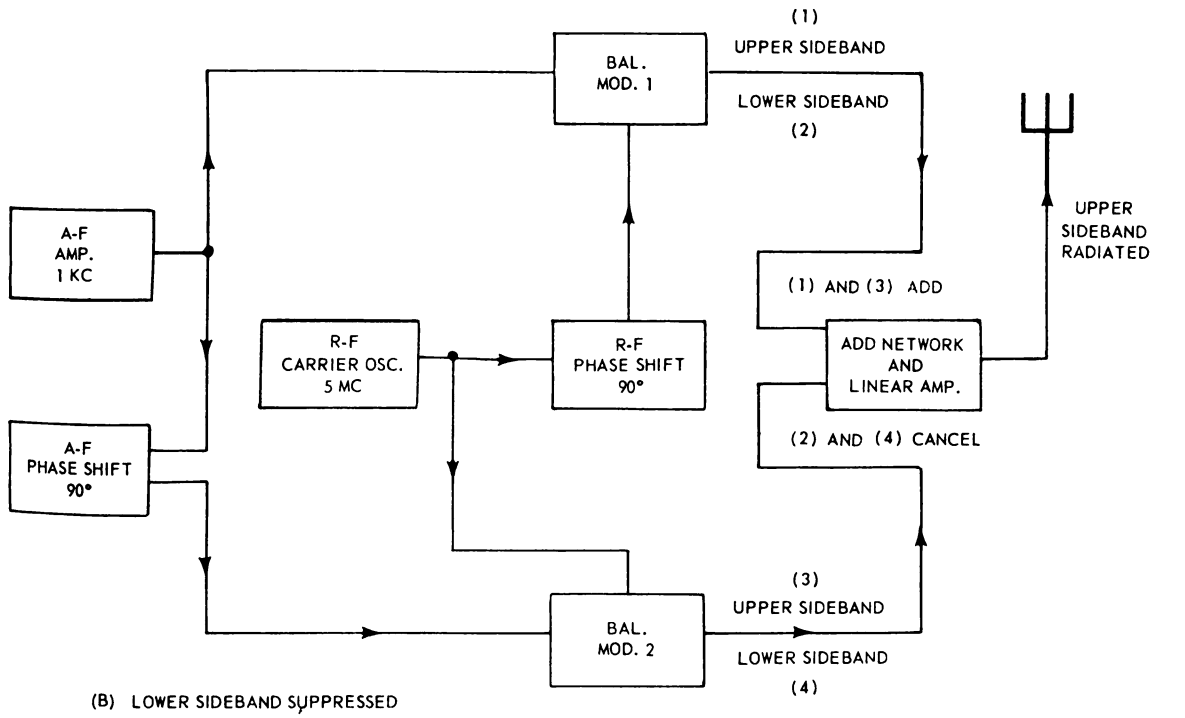


(A) UPPER SIDEBAND SUPPRESSED



(B) LOWER SIDEBAND SUPPRESSED

Figure 8-8. —Phase-shift method of SSB generation.

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the upper sideband from balanced modulator 2, and therefore these two signal components are cancelled in the add network. However, the lower sideband from balanced modulator 1 is in phase with the lower sideband from balanced modulator 2, and these components add in the add network.

As in any balanced modulator, the carrier (5 mc, in this case) is balanced out and does not appear in the output.

In figure 8-8, B, the a-f phase shifted signal is fed to balanced modulator 2, and the r-f phase-shifted signal is fed to balanced modulator 1. As indicated in the figure, the upper sideband from balanced modulator 1 combines in phase with the upper sideband from balanced modulator 2. However, the two lower sidebands combine 180° out of phase (one is advanced by 90° and one is retarded by 90°); and these components are cancelled in the add network. As in part A, the carrier is balanced out and does not appear in the output.

It is obvious that if the unwanted sideband is phased out by means of the phase-shift circuits and the balanced modulators, a narrow-band filter is unnecessary.

COMPARISON OF FILTER AND PHASE-SHIFT METHODS OF SSB GENERATION

The audio input and the linear power amplifier circuits used in each of the two methods of SSB generation are comparable. However, there are great differences in some of the other circuits.

Modulation in the filter system must take place at a low radio frequency (250 kc, for example) because the percentage of separation of sidebands at low carrier frequencies is much greater than at high carrier frequencies. The filters (electromechanical or crystal lattice) must be very selective and pass only the desired narrow band of frequencies. This will be either the lower or the upper sideband.

Because the system starts with a low generated frequency, (perhaps 250 kc) it must be heterodyned in a sufficient number of balanced modulator stages to bring it to the desired radio frequency.

In the filter system it is difficult and expensive to switch from one sideband to the other. However, this system is very stable; and once the initial adjustments are properly made, the equipment should operate with a minimum of internal adjustments.

Generation of SSB in the phase-shift system may take place in one equivalent push-pull stage at the radiated frequency, and thus the heterodyning process may be minimized. Because the unwanted sideband is phased out in the balanced modulator, the use of expensive electromechanical or crystal lattice filters is avoided.

In the phase-shift method, either of the sidebands may be used. This is accomplished by means of a simple switching arrangement. A major disadvantage of the phase-shift method is that the phase-shift networks require critical adjustments and are likely to require attention from time to time.

NARROW BAND FILTERS

The narrow band filters used after the first balanced modulator in transmitters employing the filter method of SSB generation and after the mixers in their associated receivers must have very high Q's and sharp cutoff characteristics. Mechanical or crystal lattice filters may be used. Crystal lattice filters have Q's of 10,000 to 200,000 or even higher. Mechanical filters have Q's of 2000 to 10,000 or more. Compact lattice filter units have been designed to operate up to 40 mc. However, at lower frequencies (for example, 200 kc to 250 kc) mechanical filters have proven to be compact and durable. Both types are discussed in the following paragraphs.

MECHANICAL FILTERS

Several types of mechanical (electromechanical) filters have been designed for use with SSB equipment. Fundamentally, they all operate on the same principle. Metal rods or disks are mounted in a container in such a way as to form a group of mechanical resonating elements. Energy is put into the system by means of a magnetostrictive transducer, and energy is taken out the same way. Only a relatively narrow band of frequencies can cause the resonating elements to respond. Therefore, only these frequencies can produce a signal in the output transducer.

One type of disk mechanical filter (actually, an electromechanical filter) is illustrated in a simplified form in figure 8-9. The signal is fed into the input coil, which causes the first disk to vibrate because of magnetostrictive action of the coil and the driving wire. (The

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wire is a magnetic material.) The vibrations are coupled to the remaining disks by means of the coupling wires. The vibrations of the last disk cause the output driving wire to vibrate, and by the inverse magnetostrictive effect an output signal is developed in the output coil.

The disk resonators are precisely ground to resonate at frequencies very close to the center frequency of the pass band. The width of the pass band depends on the coupling elements, the center frequency depends on the size of the resonator elements, and the selectivity depends on the number of resonant elements.

A simplified diagram of another type of mechanical filter is illustrated in figure 8-10. This filter has seven resonant sections, and two quarter-wave end sections for support. The center resonator is encircled by a "snubber" bracket, (not shown) which prevents excessive excursions under shock. The hermetically sealed housing into which the filter is mounted contains a dry inert gas. This filter vibrates with a twisting motion (torsional vibration); the twisting motion is passed from one element to another through the coupling sections.

As in the case of the disk-type filter the center frequency depends on the resonant frequency of the tuned elements, the selectivity depends upon the number of tuning elements, and the width of the pass band depends on the design of the coupling elements.

The transducers at the input and output are specially processed ferrite rods. The resonator rods are made of a specially prepared nickel

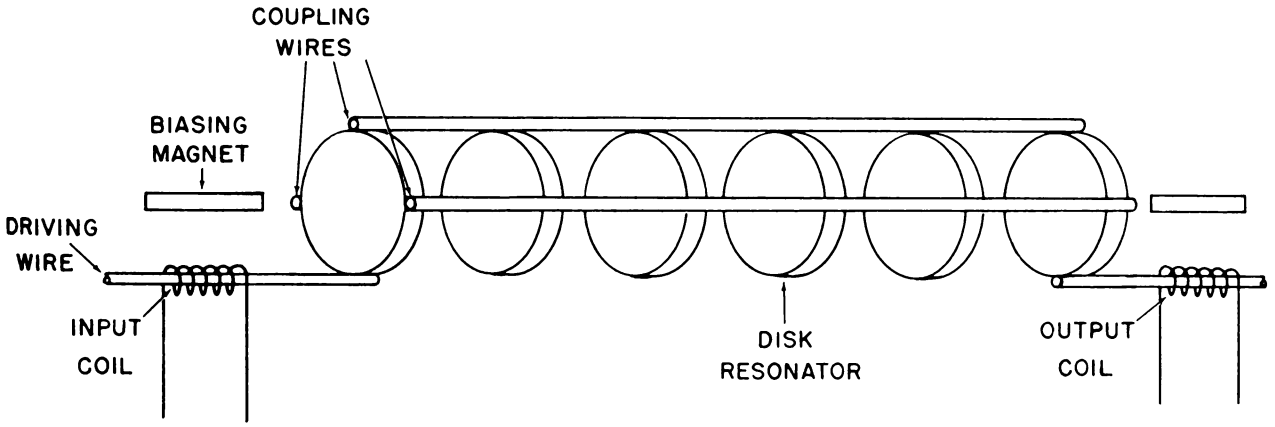

Figure 8-9. —Disk-type mechanical filter.

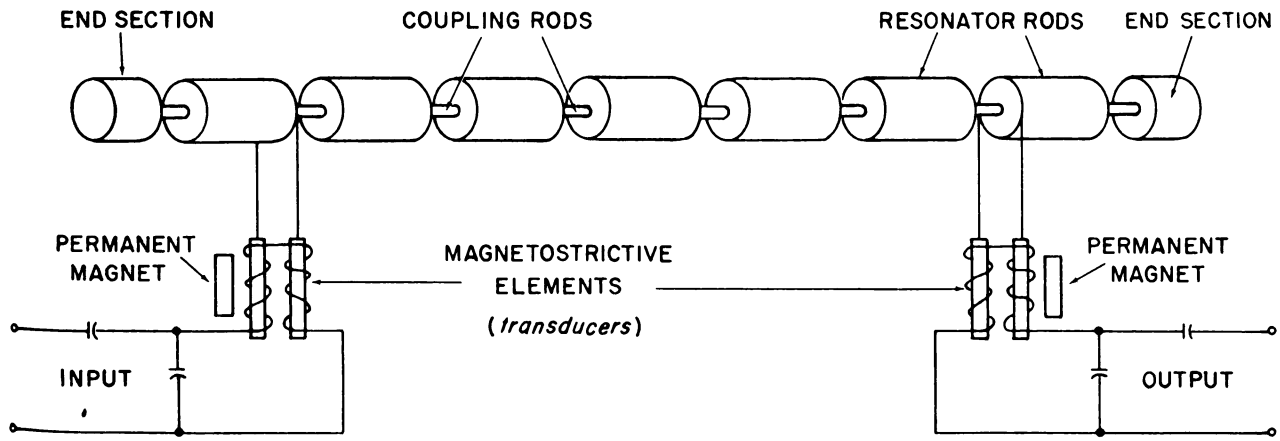

Figure 8-10. —Torsional-type mechanical filter.

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alloy heat-treated to maintain an essentially constant frequency even during wide temperature changes.

CRYSTAL LATTICE FILTERS

The equivalent electrical circuit of a quartz crystal is illustrated in figure 8-11. Capacitor C1 represents the reciprocal of the crystal stiffness—that is, compliance, which is the equivalent of capacitance in the electrical system; L represents the electrical equivalent of the crystal mass that is effective in causing mechanical vibration; and R is the electrical equivalent of internal resistance. Capacitance C2 represents the capacitance of the crystal holder plates with the crystal plate between them. Because of the very large value of the crystal Q, the equivalent circuit may be considered to be entirely reactive for most filter applications.

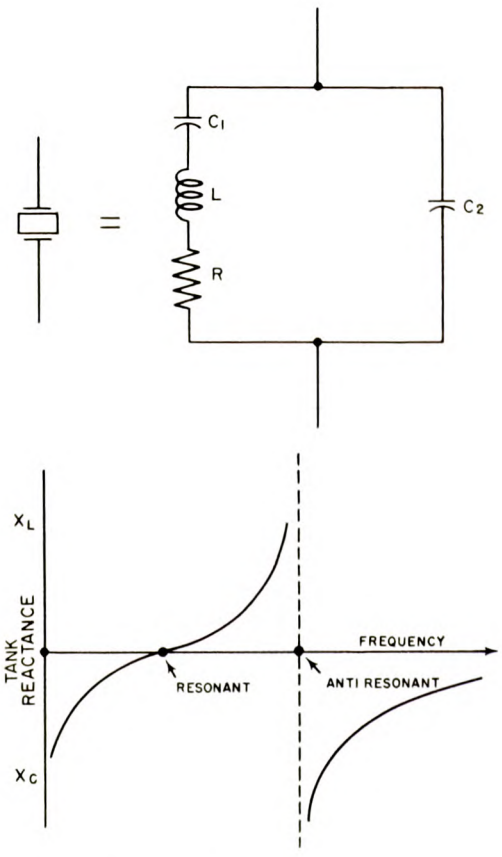

Figure 8-11. –Equivalent electrical circuit of a crystal in its holder.

As indicated on the reactance curve (fig. 8-11) series resonance occurs at the point where the reactance curve crosses the zero line. In general, the resonant (series resonant) and antiresonant (parallel resonant) points will be fairly close together. However, by the proper use of input and output reactance the points may be spread apart for filter applications.

Typical crystal lattice circuit arrangements are illustrated in figure 8-12, A. Additional elements may be added in the branches that are indicated by dashed lines.

A circuit that may be used in a SSB system is illustrated in figure 8-12, B. The parallel-connected crystals are frequency matched as close as possible; the series-connected crystals are also matched. The resonant frequency of the parallel crystals is perhaps 3 kc higher than that of the series crystals. Both the input and the output circuits are tuned to the center frequency of the pass band. In addition to the capacity of the trimmer capacitor (used for adjustment purposes) there is, of course, the distributed capacity of the crystal elements.

The crystal lattice filter is, in effect, a bridge circuit, as illustrated in simplified form in figure 8-13. Assume that the center frequency at the input is 101.5 kc and that the crystals have resonant frequencies, as shown.

At 101.5 kc, the reactances of crystals 1 and 4 will be capacitive (see reactance curve, fig. 8-11), and the reactances of crystals 2 and 3 will be inductive. The bridge will therefore be unbalanced, and a voltage will appear at the output. The same type of reasoning may be applied for other frequencies within the bandpass. For simplicity, the effects of the input and output inductancas and the distributed capacitances have been neglected.

AUDIO CIRCUITS

With the exception of the receiver first audio amplifier, the audio input and output circuits are included in the power supply chassis. For simplicity, they were not included in the simplified schematic diagram of figure 8-4. Both the audio input and output sections are treated briefly in the following paragraphs.

INPUT SECTION

A simplified schematic diagram of the audio input circuits is given in figure 8-14. When the

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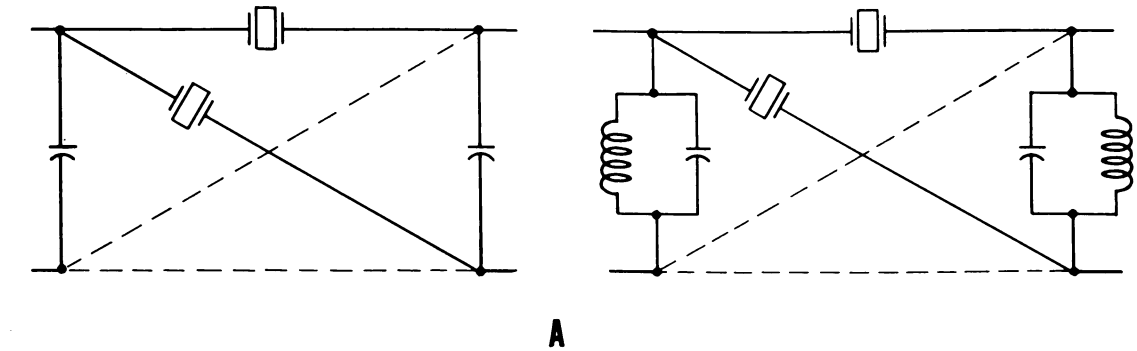
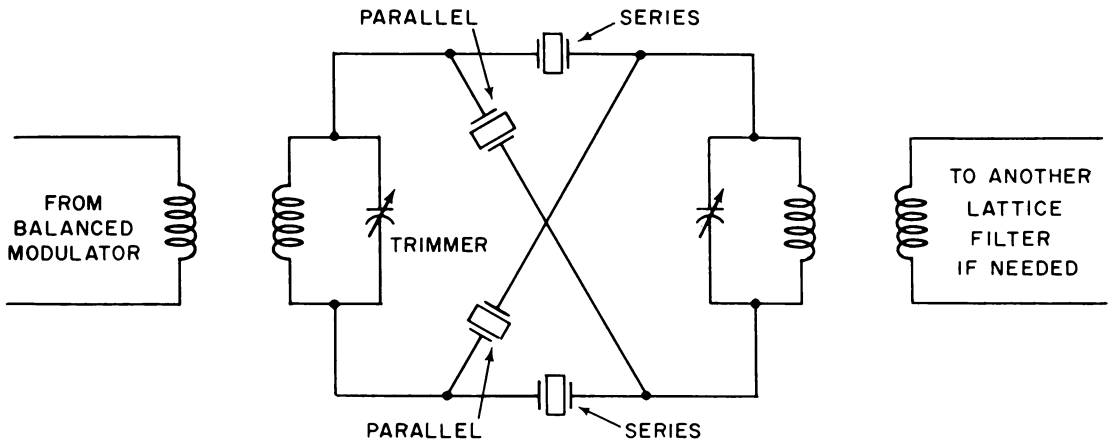


A

CIRCUIT ARRANGEMENTS



B

TYPICAL CIRCUIT THAT MAY BE USED IN A SSB SYSTEM

Figure 8-12. —Crystal lattice filter configurations.

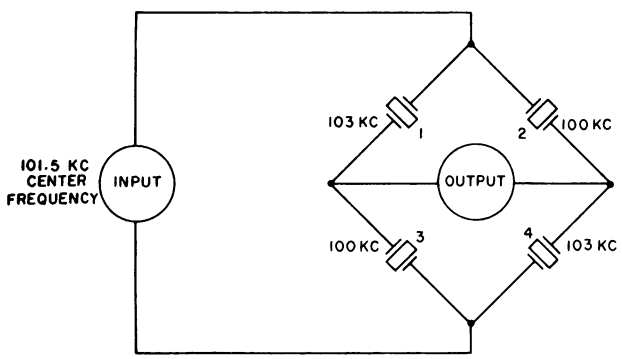


Figure 8-13. —Simplified equivalent bridge circuit.

telegraph-phone switch is in the PHONE position, the grid of the tone oscillator is grounded and the grid of the microphone amplifier is ungrounded. If one then speaks into the microphone, an output voltage is developed across R1. This voltage is fed through C1 to the grid of V1. The lower audio frequencies are attenuated in C1, and the higher audio frequencies are shunted by C2. Actually, the whole system is designed to pass from 350 to 3000 cycles.

The microphone amplifier, V1, feeds the cathode follower, V2. Tube V2, in turn, feeds the input to the 1st balanced modulator.

When the telegraph-phone switch is in the TELEGRAPH position, the microphone output is

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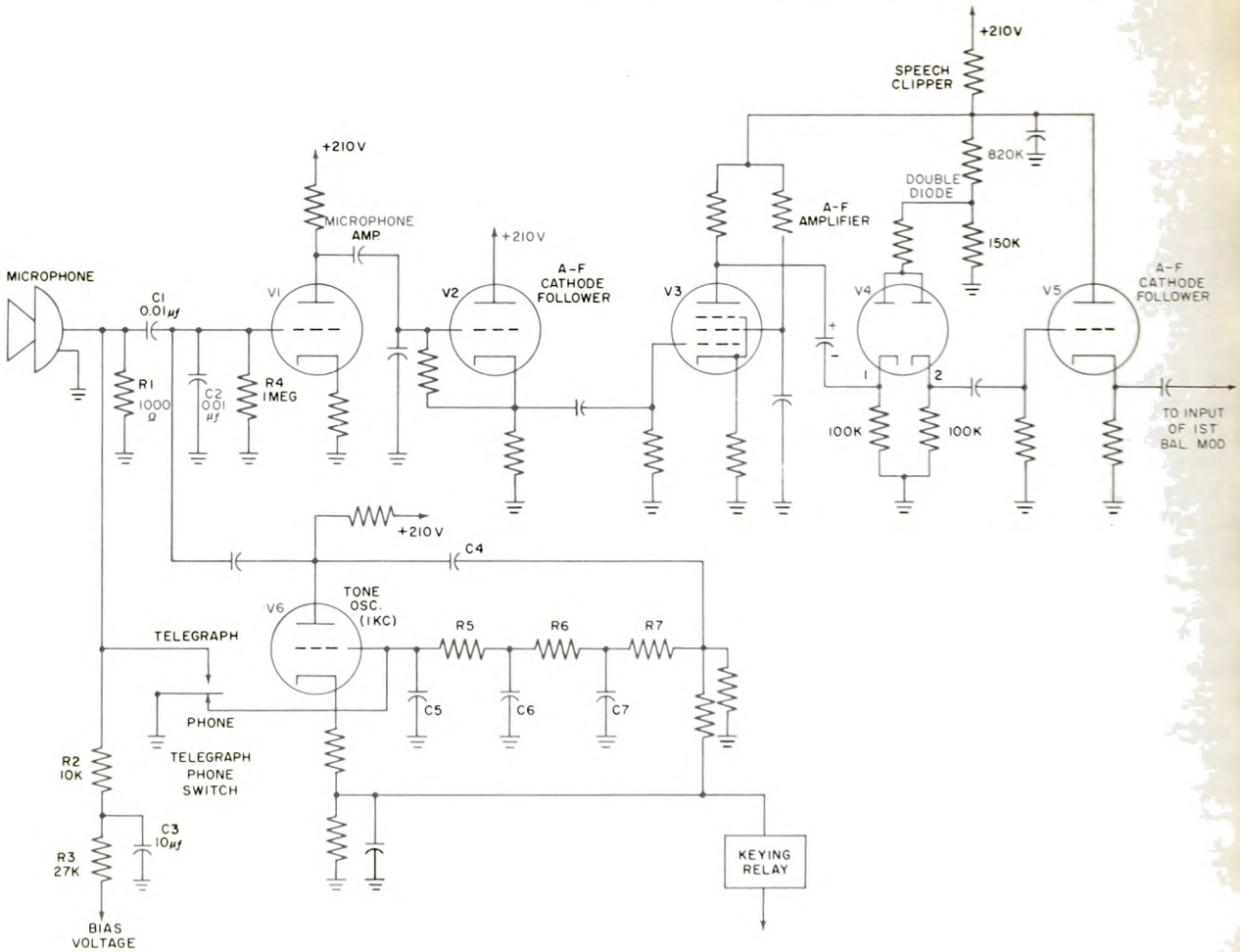

Figure 8-14. –Audio input circuits.

shunted to ground, and the tone oscillator, V6, is connected to V1.

The tone oscillator (phase-shift oscillator) provides the tone for telegraph work. A phase-shift oscillator may have any of four common circuit arrangements involving a network of resistors and capacitors or resistors and inductors coupling the plate circuit to the grid circuit. The arrangement shown in figure 8-14 is representative. In any case, the purpose of the network is to shift the phase of the signal voltage developed in the plate circuit 180° and to apply this voltage to the grid of the tube. In the circuit under consideration the network, consisting of R7C7, R6C6, and R5C5, perform the phase-shift function.

The normal amplifying action of a tube introduces a 180° phase shift between the grid voltage and the plate voltage. The phase-shift network takes the signal from the plate and introduces an additional 180° phase shift and applies the signal as a positive feedback to the grid of the tube. Sustained oscillations at one specific frequency are thus obtained.

OUTPUT SECTION

The a-f signal from the first a-f amplifier, V16 (fig. 8-4) is fed through a transformer to the input of the second a-f amplifier (not shown). The amplified signal is then fed to the grid of the a-f output tube and the output of this tube

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s applied to the speaker. Feedback is used to mprove the response.

LINEAR AMPLIFIERS

A linear amplifier develops an output that is irectly proportional to the amplitude of the nput signal. The linear amplifier used in SSB ystems must not only have low distortion but t must also provide maximum power gain.

Low distortion in the SSB system is very mportant, especially when both sidebands are adiated simultaneously and each has a dif- erent intelligence content. Distortion can cause nterference and noise effects between channels.

High-power gain in the SSB system is also very important because the modulators are operated at low-power levels. If many amplifier stages are used, the cost, size, and maintenance problems are increased.

The main problem in designing linear amplifiers is to combine low distortion and high-power gain. Normally, these two factors are not compatible.

A more detailed discussion of the linear power amplifier is included in Chapter 9 of *Basic Electronics*, NavPers 10087.

SHORE-BASED SINGLE-SIDEBAND EQUIPMENT

TRANSMITTERS

The AN/FRC-10

The transmitter portion of Radio Set AN/FRC-10 is an example of a shore-based, single-sideband transmitter. The signal is amplitude modulated, the carrier is suppressed, and the upper and lower sidebands are utilized as twin channels. It is designed to satisfy the needs of long-range, point-to-point communication in the 4- to 23-mc frequency range. Conventional double-sideband a-m facilities are provided in this transmitter for use when a single-sideband receiver is not available.

Any one of ten preset operating frequencies may be selected by means of pushbuttons. The peak envelope power output for single sideband is 4 kw; for double sideband, the carrier output is 1 kw.

Two AN/FRC-10 terminal sets, including the necessary carrier terminals and companion receivers, make up a two-way system that can handle communications between points thousands of miles apart.

A simplified block diagram of two communications terminals is given in figure 8-15. By using proper terminal equipment, the transmitter can accept two groups (group A and group B) of signals, each 100 to 6000 cycles wide. Group A becomes the intelligence of one sideband, and group B becomes the intelligence of the other. Group A may consist of six teletype channels, and group B may consist of a band of radio-telephone signals. Group B could be divided into two voice circuits or one voice and one facsimile circuit. Other combinations could, of course, be used. The teletype and voice frequency signals fed to group A and group B are multiplexed signals and are described in the next chapter.

A simplified block diagram of the transmitter portion of the AN/FRC-10 system is given in figure 8-16.

Group A and group B frequencies, each having a bandwidth of 100 to 6000 cps are fed to the pad and relay circuits. The pads equalize the signal levels in the two groups, and the relays are used to interchange group A and group B inputs to the 1-f modulators for frequencies above 10 mc. That is below 10 mc, channel A is the lower sideband and above 10 mc, channel A is the upper sideband. Another relay connects group A signals to the DSB modulator (when this is desired) and opens the group B circuits.

One signal group is fed to 1-f balanced modulator 1, and the other is fed to 1-f balanced modulator 2. The 100-kc local oscillator also feeds balanced modulators 1 and 2. The carrier is reduced about 40 db below the reference level in the output of the 1-f balanced modulators. The two sidebands (94 to 99.9 kc and 100.1 to 106 kc) appear in the output of both modulators. The upper sideband is passed by the upper filter (1), and the lower sideband is passed by the lower filter (2).

The two sideband frequencies from the filters are fed through a hybrid coil (see lower portion of figure 8-16) to the input of the load control amplifier. The hybrid coil delivers the 100.1 to 106-kc band and the 94- to 99.9-kc band to the grids of the load control amplifier tubes by combining the outputs of filters 1 and 2 in series opposition from one pair of windings to the grids, and adding the outputs of filters 1 and 2 from another pair of windings in series across an isolated common load resistor. The input to the load control amplifier is centered about a greatly suppressed reference frequency of 100 kc. The bias of this amplifier (and its

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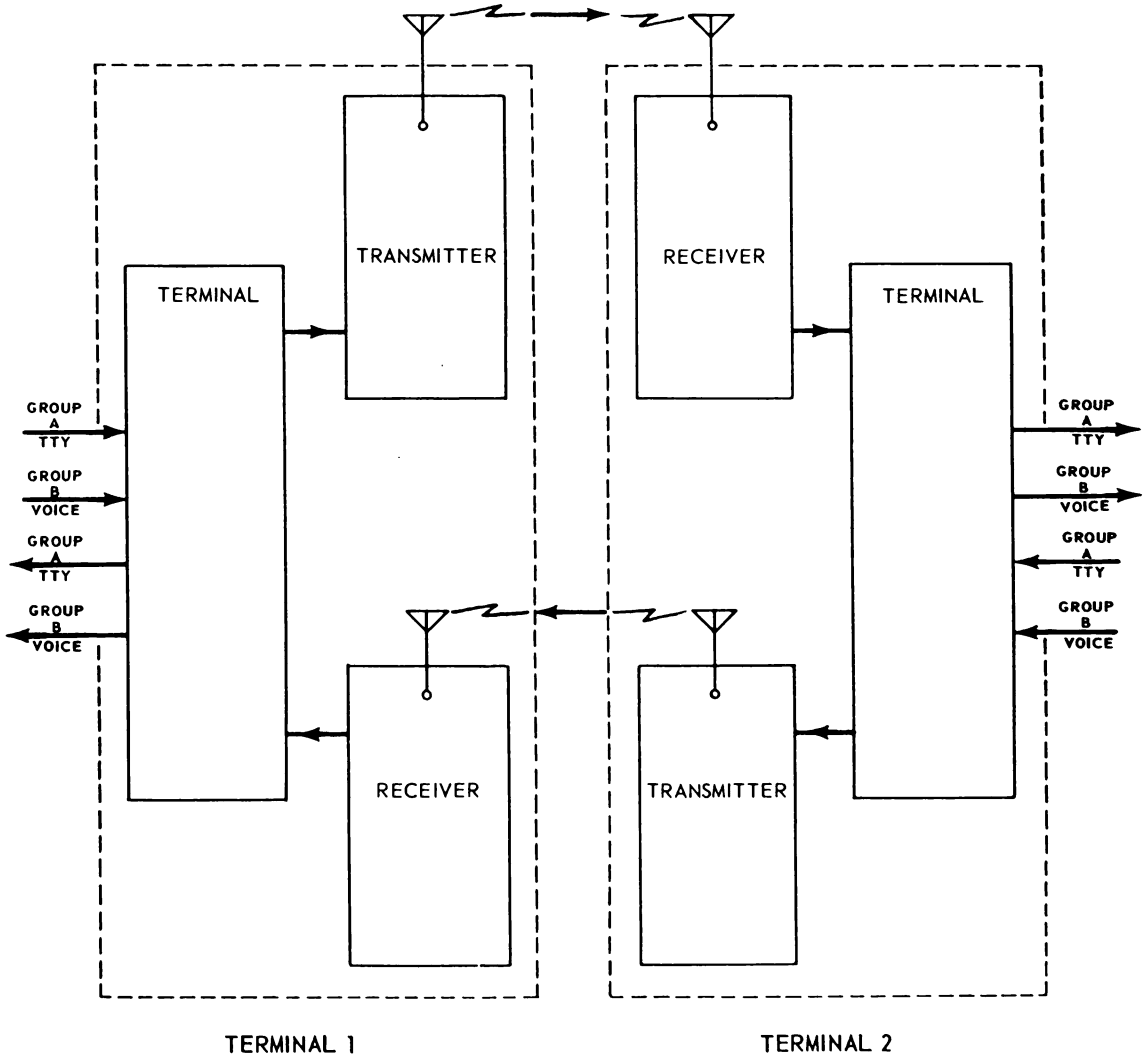

Figure 8-15. -Block diagram of the AN/FRC-10 system.

gain) is controlled by a voltage obtained from the load control and output indicator rectifier. The input to the load control is obtained from the h-f amplifier. The purpose of the load control circuit is to keep the transmitter peak voice loading at a given level.

The l-f combining amplifier may be supplied with either of two inputs. It may be fed a conventional carrier double-sideband signal from the DSB modulator or a resupplied carrier to be combined with the output of the load control amplifier. The position of the carrier resupply switch determines the input to the l-f combining amplifier.

As in the case of the l-f balanced modulator, the m-f balanced modulator has two inputs. One input (100 kc) is from the l-f combining amplifier and the other is from the 2700-kc oscillator. The 2700-kc signal is greatly suppressed; however, the modulation products include a 2600-kc carrier and a pair of associated sidebands and a 2800-kc carrier and a pair of associated sidebands. Only the 2800-kc carrier and its sidebands are passed by the m-f amplifier.

The h-f balanced modulator also has two inputs. One is from the m-f amplifier and the other is from the 6.8- to 20.2-mc generator. The output contains two pairs of sidebands: one is 2.8 mc below the final conversion frequency and the other is 2.8 mc above the final conversion frequency. The h-f power amplifier passes

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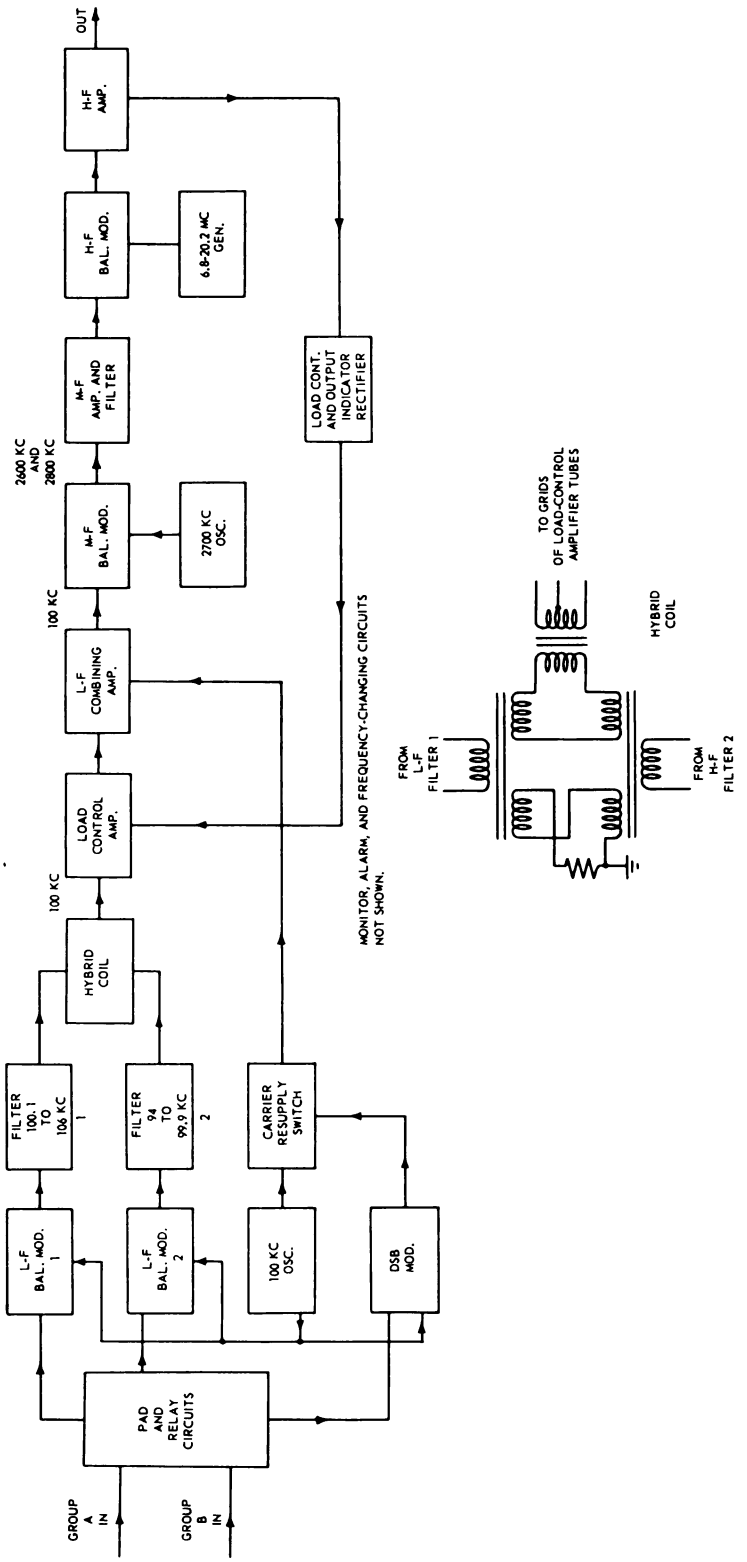

Figure 8-16.—Simplified block diagram of the AN/FRC-10 transmitter.

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only one pair of sidebands appearing at the output of the h-f balance modulator. The other sideband pair is rejected. These frequencies are further amplified in linear amplifiers, and radiated.

The TEF

The TEF transmitter is designed to be used for twin-channel, single-sideband or single-channel, double-sideband transmission. A switching arrangement permits the operator to select the desired type of transmission. When double sideband is used, the carrier frequency may be "wobbled" plus and minus 250 cycles from the mean value, if additional privacy is desired.

A simplified block diagram of the TEF transmitter is shown in figure 8-17. When used as a single-sideband transmitter, two channels may be transmitted. The two channels are present in the output as two independent sidebands, one on each side of the carrier frequency. For single-sideband operation the amplitude of the carrier is adjustable between the full-power output capability of the transmitter and a value about 30 db below either sideband. When only one single-sideband channel is being transmitted, there is present only the reduced carrier and one sideband.

The transmitter is designed to operate on any frequency between 4500 kc and 22,000 kc. It is capable of delivering a peak envelope power of 2 kw (conventional double-sideband operation) to an open-wire transmission line having an impedance of 400 to 800 ohms, or to a balanced coaxial transmission line having an impedance of about 200 ohms.

The transmitter circuits are so designed that the carrier frequency may be shifted to any one of six preselected frequencies. Three modulation steps are employed. A 625-kc crystal oscillator is the source from which conversion frequencies for the first two stages of modulation are derived; a h-f crystal oscillator, feeding through a harmonic generator, supplies the third balanced modulator.

A 125-kc multivibrator, which is held in step by the 625-kc crystal oscillator, supplies the first modulator. The 625-kc crystal oscillator also feeds the 2500-kc harmonic generator; this generator operates at the fourth harmonic of the crystal oscillator.

The 125-kc frequency serves as a conversion frequency for the first step of modulation, which occurs in modulators 1A and 1B for channels A and B, respectively. The upper sideband is selected from the output of modulator 1A, and the lower sideband is selected from the output of modulator 1B by crystal filter A and crystal filter B, respectively.

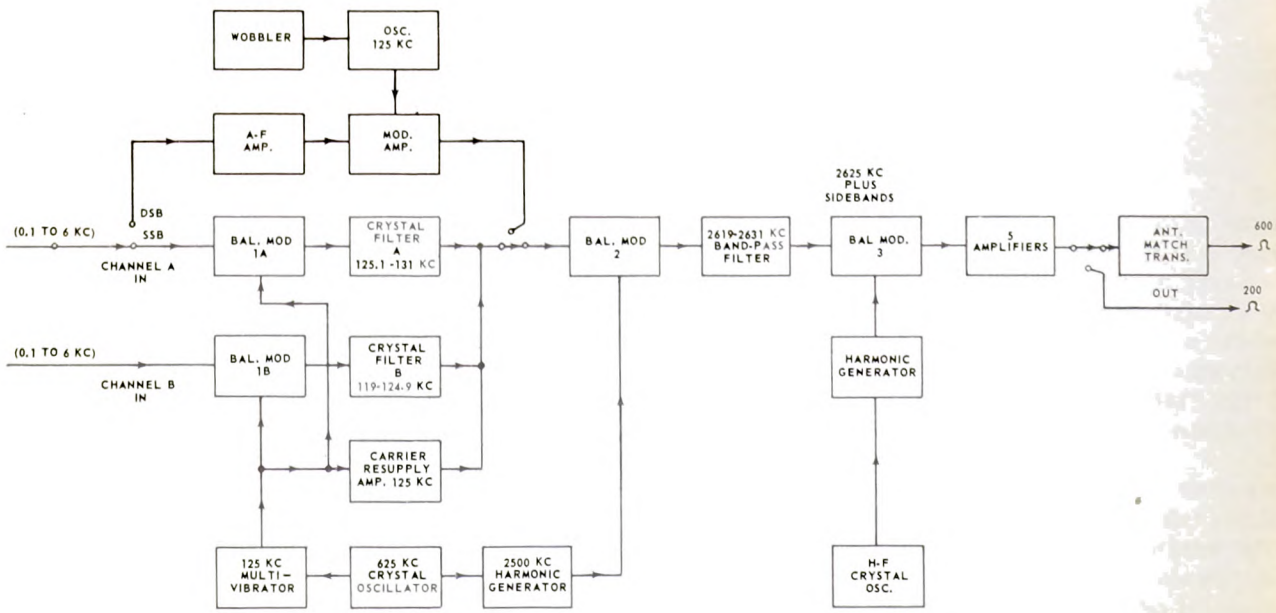

Figure 8-17. –Simplified block diagram of TEF transmitter.

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The outputs of the two crystal filters are combined and impressed upon the input of modulator 2, along with the output of the carrier resupply amplifier. The 125-kc output of the carrier resupply amplifier is adjustable between the full output power and a value of about 30 db below either sideband.

The 2500-kc (from the 2500-kc harmonic generator) serves as a conversion frequency for the second step of modulation in modulator 2. The output of modulator 2 is fed to the 2619- to 2631-kc band-pass filter, which suppresses the 2500-kc conversion frequency (any portion not suppressed in the balanced modulator) and the lower sideband resulting from the action of modulator 2. The upper sideband contains a 2625-kc carrier frequency and two independent sidebands; the upper of which contains frequencies dependent upon the input of channel A, and the lower of which contains frequencies dependent upon the input of channel B.

These frequencies (the 2625-kc carrier and the two independent sidebands) are fed to modulator 3. The conversion frequency for modulator 3 is derived from the harmonic generator, which is driven by the high-frequency crystal oscillator. The frequency of the high-frequency crystal oscillator is determined by the required final carrier frequency. If the assigned frequency is f, the high-frequency crystal oscillator and harmonic generator must be so adjusted that the output frequency of the latter is either f + 2625 kc or f - 2625 kc.

When f + 2625 kc is used (for assigned carrier frequencies below 10,000 kc), the lower sideband produced by modulator 3 is transmitted. When f - 2625 kc is used (for assigned carrier frequencies above 10,000 kc), the upper sideband is transmitted. Hence, to shift the carrier frequency, it is only necessary to change the conversion frequency supplied to modulator 3 and retune the transmitter.

A series of five amplifiers follows modulator 3. These amplifiers increase the power level of the desired sideband obtained from the third step of modulation to a peak envelope power of 2 kw. The output of amplifier 5 may be switched to either an antenna matching transformer or a balanced coaxial transmission line. The output circuit of amplifier 5 is designed to operate into a balanced load of about 200-ohm impedance. The antenna matching transformer permits matching the amplifier output to a balanced open-wire transmission line having an impedance of 400 to 800 ohms.

Although the preceding description applies to single-sideband transmission, the principle of operation is essentially the same for double-sideband transmission. When double-sideband transmission is used, the channel A input is switched from modulator 1A to the audio-frequency amplifier. The 125-kc oscillator frequency is modulated by the audio frequency in the modulated amplifier and fed to modulator 2. When the equipment is being used for double-sideband transmission, the 125-kc multivibrator is disabled to prevent operation of the single-sideband components.

Single-sideband transmissions may be monitored by a monitor panel (not included in the block diagram). A small portion of the output of amplifier 5 is demodulated by means of two stages of demodulation and fed to the monitor output. Double-sideband transmissions may be monitored by the output indicator circuit provided in amplifier 5.

RECEIVER

BLOCK DIAGRAM

An example of a shore-based single sideband radio receiving set is the AN/FRR-10. This set is a double-conversion, superheterodyne-type diversity receiver contained in three rack cabinets. The complete receiving equipment consists of a group of two radio receivers together with the switching and comparing circuits necessary to receive the following types of signals in single channel or space diversity operation:

1. Double-sideband radio telephone
2. Single-sideband radio telephone and teletype
3. Single-sideband radio telephone or teletype with carrier suppression up to 26 decibels.

The receiver covers the tuning range of 2.0 to 32.0 mc in four tuning bands. The three bays that comprise the AN/FRR-10 are shown in outline form in figure 8-18.

A simplified block diagram of the AN/FRR-10 is given in figure 8-19. The control panel provides a means for selecting the frequency range to be used and the method of first frequency conversion (using the oscillator within the first converter or one external to it).

The r-f amplifier-first converter amplifies the incoming r-f signal and heterodynes this signal to an output frequency of 1750 kc.

The second frequency converter converts the 1750-kc intermediate frequency obtained from

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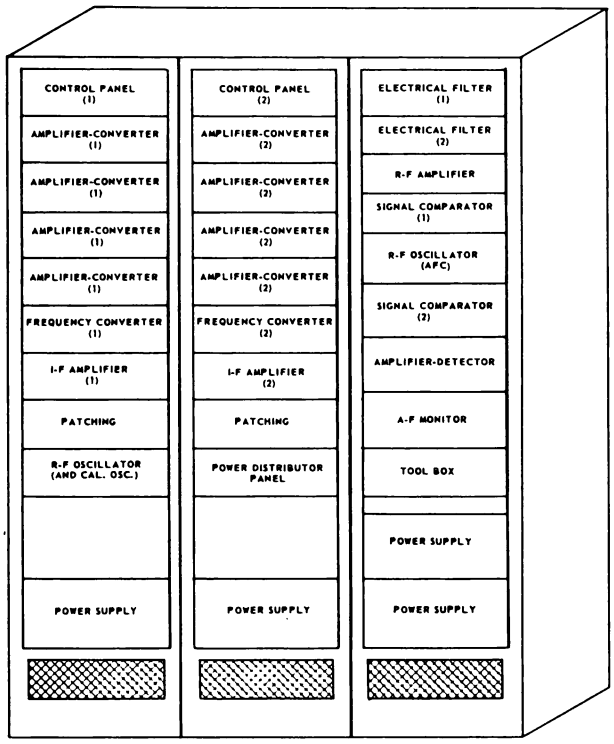

Figure 8-18. –The three bays that comprise the AN/FRR-10 receiver.

the r-f units to the 50-kc intermediate frequency of the i-f amplifier units, thus sharpening the selectivity.

The electrical filter assembly determines the final bandwidth and i-f selectivity before demodulation.

The r-f amplifier amplifies and separates the carriers from two separate composite signals fed to it from the i-f amplifiers. In addition, it supplies the crystal-controlled, 50-kc reference for automatic frequency control and demodulation.

In diversity operation the signal comparator (1) amplifies and selects the stronger of the two 50-kc carriers from the r-f amplifier and supplies two types of AGC voltage and a carrier operated-noice-silencer (CONS) voltage to other units.

The AFC oscillator provides automatic frequency control of the equipment. A motor-tuned oscillator is controlled by the output of a circuit that compares the frequency of the 50-kc crystal oscillator (contained in the r-f amplifier) and that of the incoming carrier. The motor-tuned oscillator is used as the controlling oscillator for AFC.

Signal comparator (2) amplifies and selects the stronger of two sidebands from receiver 1 and receiver 2 operating in diversity. Switches in this unit allow the comparing circuits to be bypassed for single-channel operation.

The amplifier-detector provides the demodulation and audio amplification for the equipment. It provides audio to four outgoing lines to the audio-frequency monitor.

The audio-frequency monitor provides a means of monitoring the outgoing signals by use of a self-contained loudspeaker, headphones, or a panel-mounted, line-level meter.

FUNCTIONAL DESCRIPTION OF UNITS

A detailed description of each circuit in the AN/FRR-10 will not be undertaken in this chapter. Instead, a functional description of the various units will be given. Circuit diagrams of the various units are given in the instruction book, NavShips 92114; the theory of operation of the various circuits and components is also included in this publication.

Refer to figure 8-19 as you read the following paragraphs.

The r-f amplifier first converters (4 units for receiver 1 and 4 units for receiver 2) amplify the incoming r-f signal and heterodyne it to an output intermediate frequency of 1750 kc. The frequency range of the converters for each receiver is 2 to 32 mc. One unit in each receiver covers the 2- to 4-mc range, another the 4- to 8-mc range, another the 8- to 16-mc range, and the final one the 16- to 32-mc range.

These units (fig. 8-20) comprise three stages of r-f amplification, a converter stage, a h-f oscillator stage, an AGC bias tube, a local r-f amplifier, and a remote r-f amplifier. The converter may be fed from the local h-f oscillator contained in the unit, or from another amplifier-converter unit via the remote r-f amplifier contained in the unit.

The local h-f oscillator can also be connected to another amplifier-converter unit to provide common oscillator excitation for two units in space diversity operation (see Chapters 1 and 13 of *ET3*, NavPers 10188-A). Automatic gain control voltage is connected into the unit from signal comparator (1).

For calibration purposes a crystal calibrator, located in the r-f oscillator (and CAL osc) unit, is connected to the unit. Unmodulated marker frequencies in multiples of 200 kc are available.

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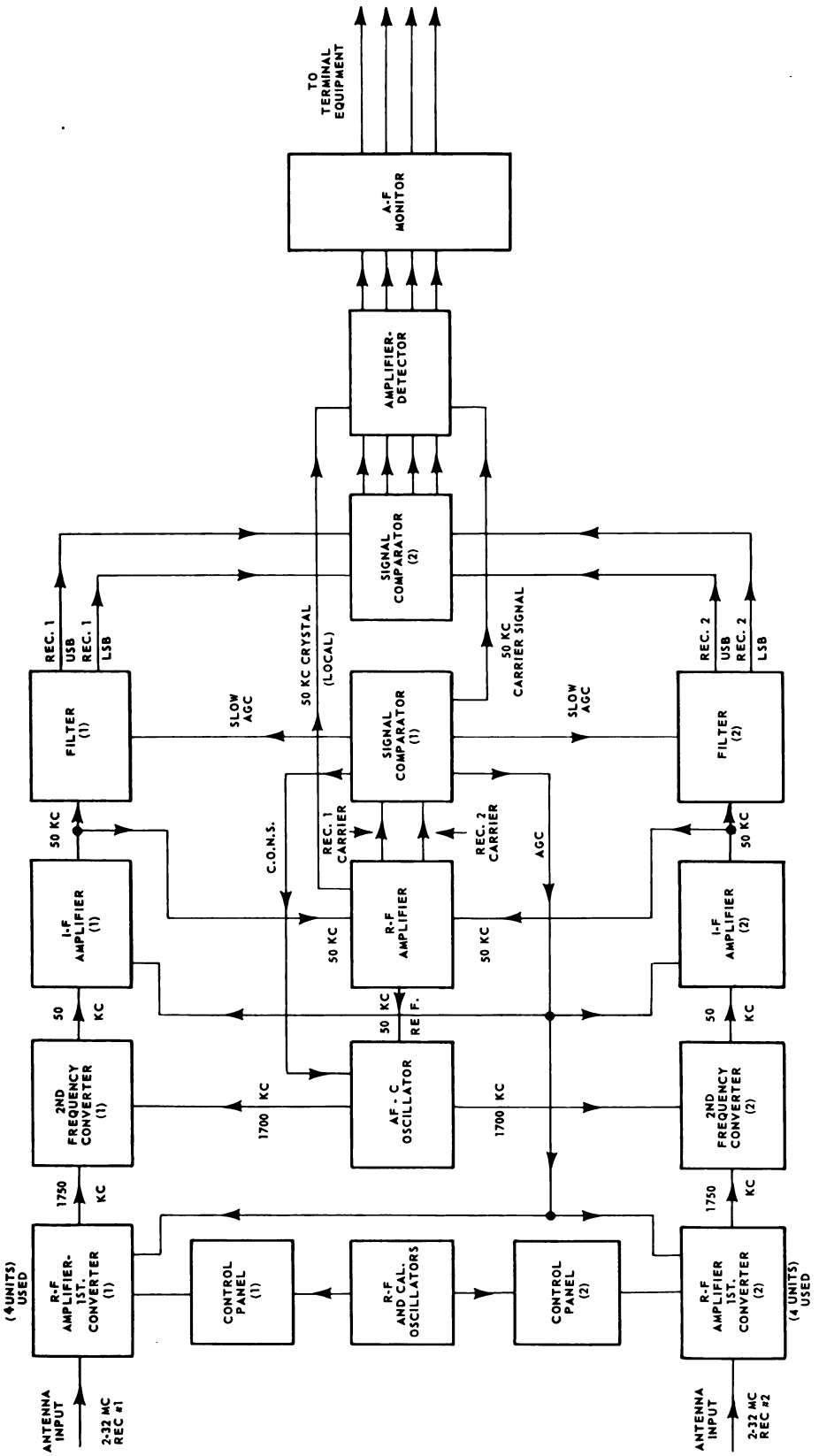

Figure 8-19. -Simplified block diagram of the AN/FRR-10 set.

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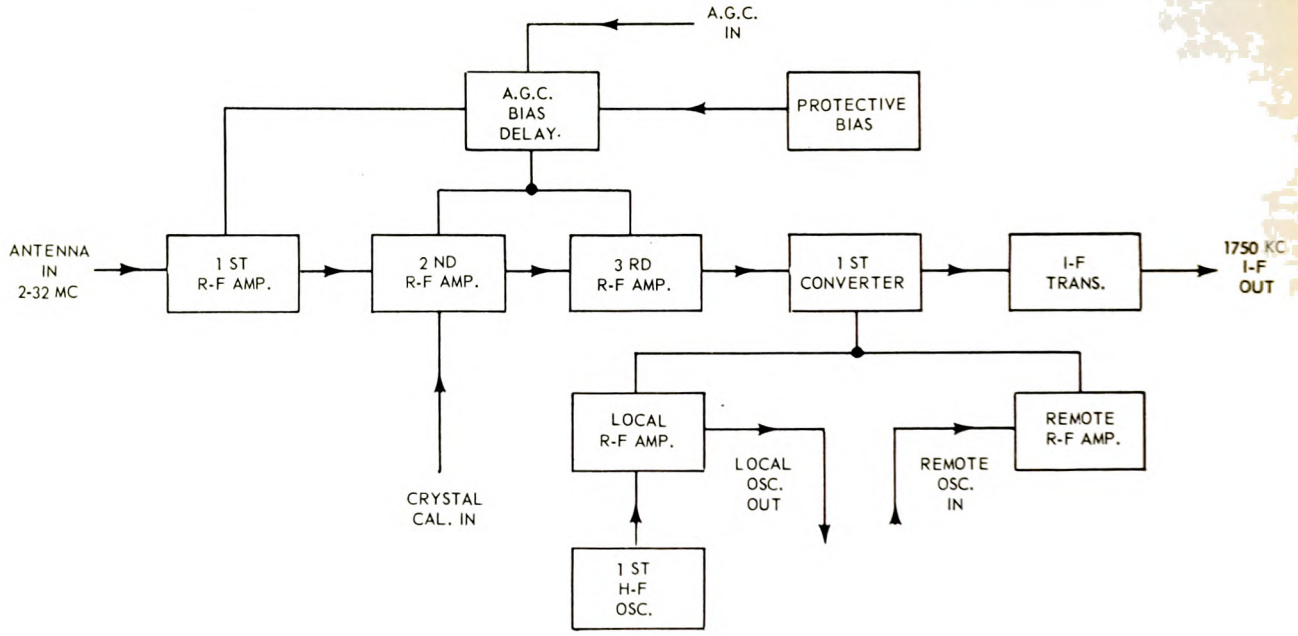

Figure 8-20. –Block diagram of an r-f amplifier and first converter.

In this receiver the first conversion oscillator operates above the incoming frequency on all bands. For example, when the incoming frequency is 2 mc (2000 kc), the first oscillator frequency is 3750 kc. If the signal is amplitude modulated with a 10-kc voltage, the upper sideband of the incoming signal will then be 2010 kc; this figure subtracted from the first local oscillator frequency of 3750 kc results in an output of 1740 kc from the first converter. The lower sideband of the incoming signal will be 1990 kc; and this figure subtracted from the first local oscillator frequency of 3750 kc results in an output of 1760 kc from the first converter. Therefore, any intelligence contained in the upper sideband of the incoming signal is now contained in the lower sideband. Likewise, any intelligence contained in the lower sideband of the incoming signal is now contained in the upper sideband.

No sideband frequency inversion occurs in the second converter because its oscillator frequency (1700 kc) is below the frequency (1750 kc) of the incoming signal.

The second frequency converter provides an intermediate step in lowering the 1750-kc intermediate frequency obtained from the first converter to the 50-kc needed for the i-f amplifier. This first conversion having a higher i-f reduces the undesired image response, and the second conversion having a lower i-f provides the required selectivity in the associated i-f amplifier. A block diagram of the second frequency converter is given in figure 8-21.

Four i-f input channels are provided in this unit to accept the 1750-kc outputs of the four amplifier converter units in the same receiver bay. The desired channel is selected by means of the band switch on the control panel.

The 1750-kc, i-f input signal is converted to an output signal of 50 kc by a converter and self-contained 1700 kc crystal or variable oscillator. The self-contained oscillators can be disabled, and the AFC oscillator can be used as a source of excitation for the converter.

Either the self-contained 1700-kc, crystal-controlled oscillator output or the variable oscillator output is available for use in other bays. The selector switch in this unit enables the operator to choose the desired oscillator.

The i-f amplifier (fig. 8-22) provides the major portion of the receiver gain and selectivity; the i-f is 50 kc. Three stages of amplification and a cathode-follower output provide an over-all gain of 80 db. Three band-pass filters (8 kc, 12 kc, and 18 kc) located in this unit provide the required selectivity for

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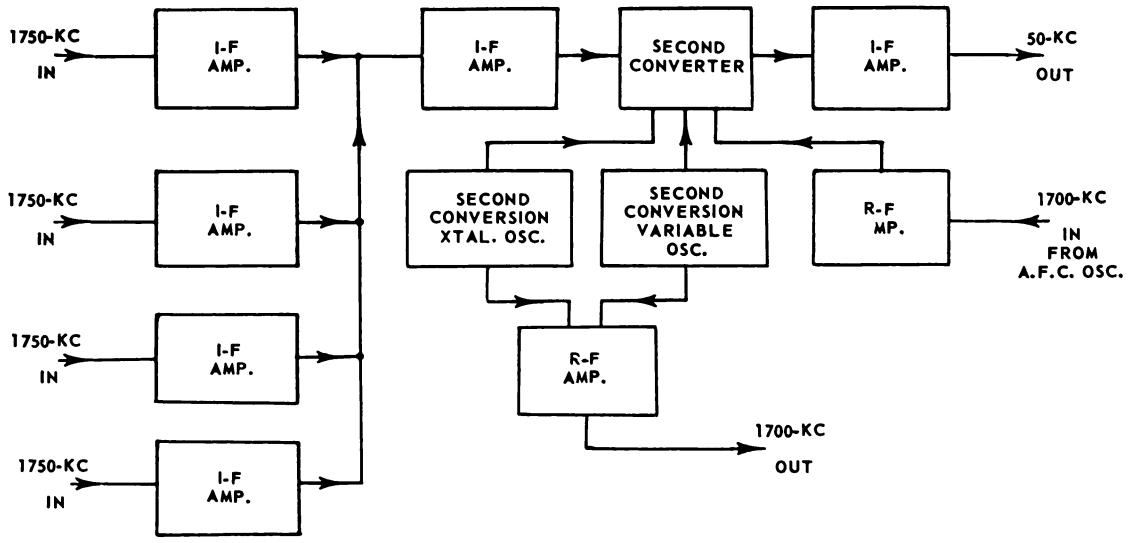

Figure 8-21. —Block diagram of a second frequency converter.

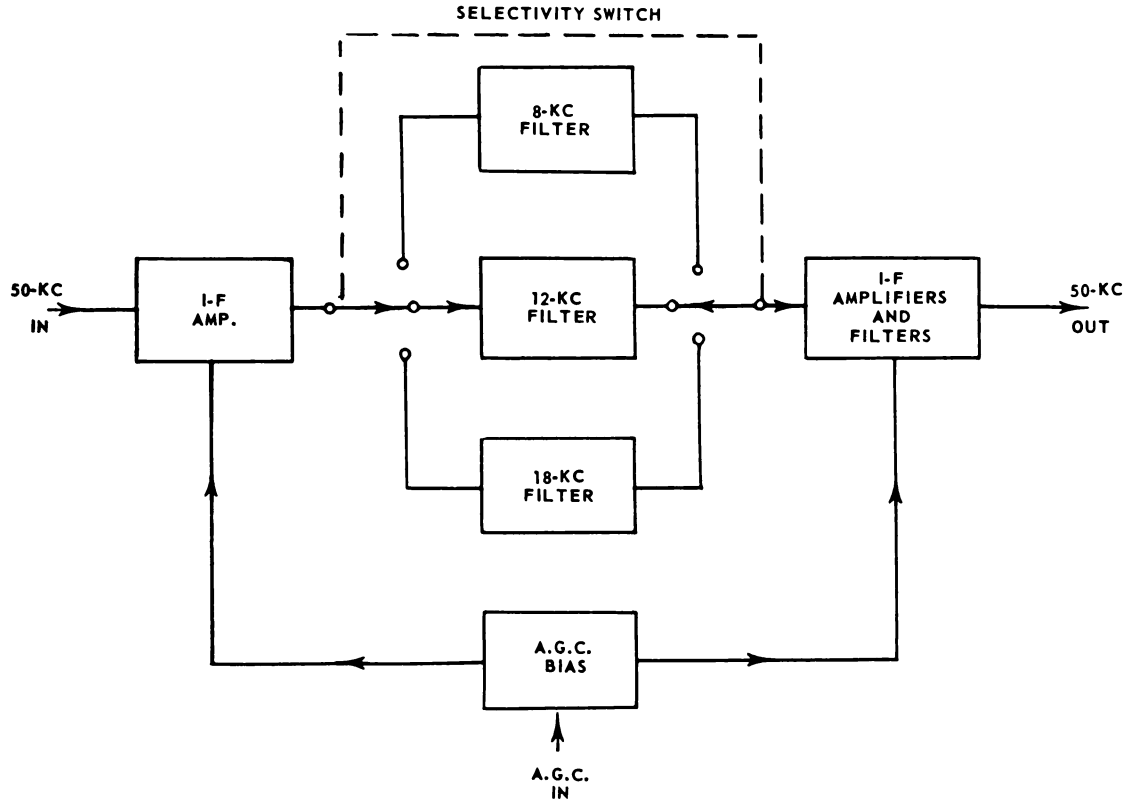

Figure 8-22. —Block diagram of i-f amplifier.

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the received signal. An AGC bias stage is incorporated in this unit as an additional protection for the tubes in case of failure of the incoming AGC.

Each filter employs seven tubes and the necessary filters to separate the sidebands from the carrier. A simplified block diagram of a filter is given in figure 8-23. The filter used in the lower sideband channel allows the passage of signals within the frequency range of 50.2 to 56 kc. (Although this appears to be the upper sideband, it actually contains the lower sideband intelligence because a frequency inversion took place in the first converter.) The filter used in the upper sideband channel allows the passage of signals within the frequency range of 44 to 49.8 kc.

Sufficient amplification is provided to overcome the filter losses. A cathode follower output provides the correct output impedance.

Signal comparator (2) is used to amplify the upper and lower sidebands from two receivers operating in diversity reception and to select the stronger of the two signals.

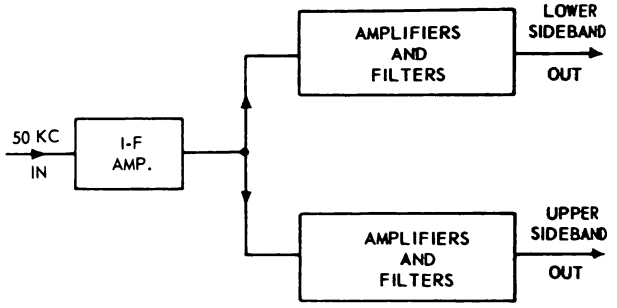

Figure 8-23. –Block diagram of filter.

The functional block diagram (fig. 8-24) will show the paths of the signals through the signal comparator. The signals from filters (1) and (2) in figure 8-24 are amplified and fed to the differential detectors, which select the stronger of the two channels. The selected upper sideband and the selected lower sideband are further amplified by gated stages, which are controlled by d-c amplifiers and flip-flops. Cathode followers are used in the outputs.

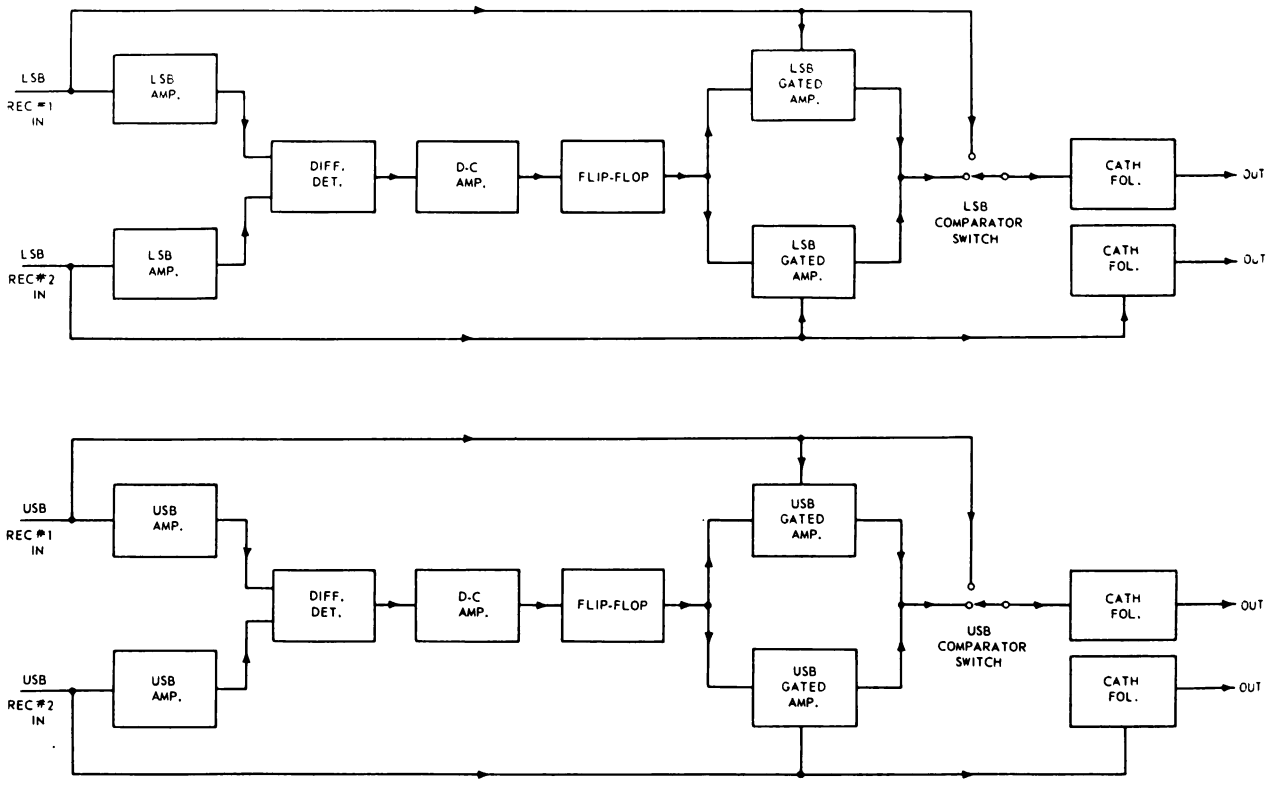

Figure 8-24. –Block diagram of signal comparator (2).

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When diversity operation is not desired, both sideband signals may be connected directly from the input of the comparator to the cathode followers by means of switches. The four output sideband signals from the comparator are connected to the amplifier detector.

The amplifier detection block diagram is given in figure 8-25. The purpose of the amplifier-detector is to demodulate and amplify any or all of the four sidebands connected to it from the signal comparator. Heterodyne detection is used to reinject the 50-kc carrier (which was suppressed at the transmitter) and thereby obtains the difference frequency between the sideband and the carrier as the detected signal. The beat frequency is supplied by either the 50-kc signal carrier or by the crystal contained in the r-f amplifier. The desired signal source is chosen by means of the carrier selector switch.

Four identical channels are contained within the unit, each channel providing independent detector and amplifier circuits. The outputs of the amplifiers are connected to the four outgoing lines via a line interchange switch. This switch makes possible the connection of any of the incoming signals to any of the four outgoing lines to the a-f monitor.

The audio-frequency monitor (not illustrated by block diagram) provides a means of monitoring any one of the four signals appearing at the amplifier detector. Two switches are provided on the front panel so that the speaker with its associated amplifier, and/or the line level meter may be connected to the desired line.

The r-f oscillator (in same unit with calibration oscillators) utilizes 16 crystals, eight for oscillator A and eight for oscillator B, as indicated in the simplified block diagram of figure 8-26, to provide heterodyne voltages for crystal-controlled operation of one or more amplifier-converter units.

Two front-panel mounted crystal switches are used to select the crystal necessary to produce the heterodyne voltage required for the reception of the chosen frequency channel. Tuning meters are provided to indicate the magnitude of each oscillator output. Individual switches permit the simultaneous operation of both oscillators.

The 200-kc oscillator and harmonic generator provide marker frequencies for injection into the amplifier-converter units for calibration checks.

The AFC oscillator (fig. 8-27) provides automatic frequency control for the receiving equipment. A reference voltage from a 50-kc, crystal-controlled oscillator in the r-f amplifier is mixed with the voltage created by the incoming carrier. Any difference in frequency existing between the two voltages actuates the capacitor drive motor, which, in turn, rotates the oscillator tuning

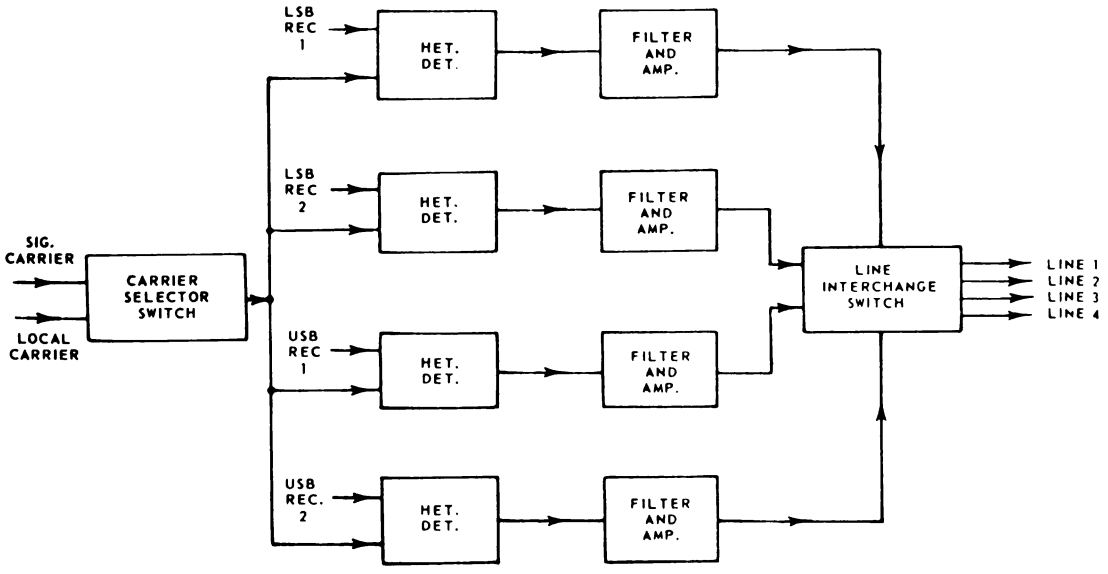

Figure 8-25. —Block diagram of amplifier detector.

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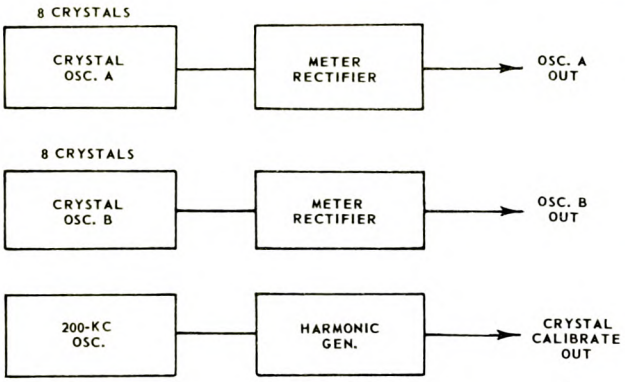

Figure 8-26. –Block diagram of r-f oscillator and calibrate oscillator.

capacitor. This causes a frequency variation in the 1700-kc oscillator in this unit. The oscillator frequency from this unit is fed to the second frequency converter. The net effect of this arrangement is to cause the voltage created by the incoming carrier to the r-f amplifier to be exactly the same frequency as the reference voltage. When this occurs, the motor comes to rest and will remain passive until such a time as frequency correction is again required.

A carrier-operated-noise-silencer (CONS) amplifier is included in the unit. Its function is to hold the motor locked in the absence of an incoming signal.

The r-f amplifier contains two separate r-f amplifiers and a crystal-controlled oscillator,

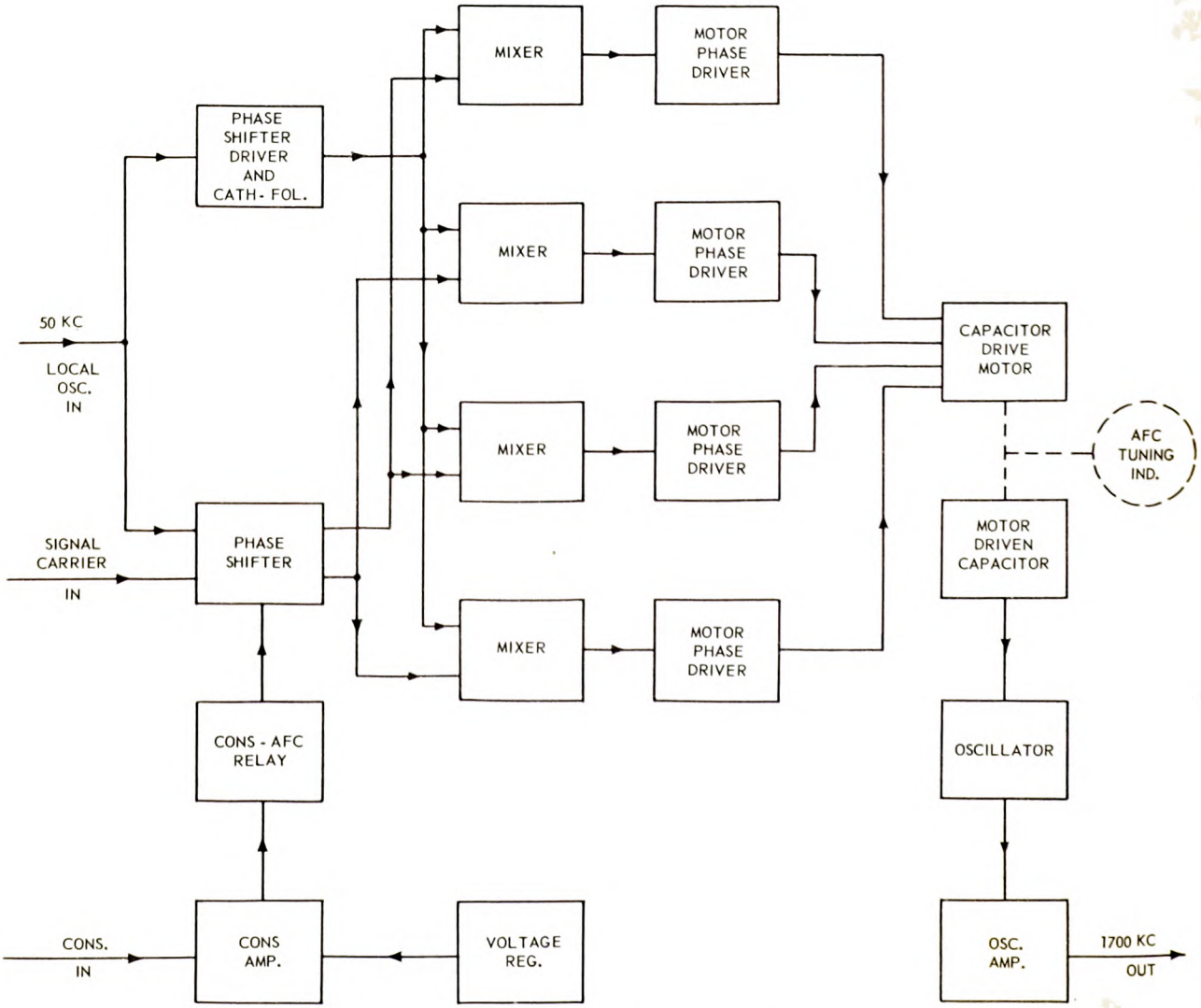

Figure 8-27. –Block diagram of AFC oscillator.

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as indicated in figure 8-28. Each r-f amplifier contains a sharp filter to separate the carrier from the sidebands of one receiving channel. Signals are fed to the unit from one or both of the i-f amplifiers. Each of these filters has a band pass of 25 cycles 6 db down from resonance. The carrier frequencies thus passed by the filters are used only for control purposes in the AFC and AGC circuits.

The crystal oscillator provides a 50-kc local signal for use with the automatic frequency control of the r-f oscillator; it is also used as a heterodyne voltage in the amplifier detector.

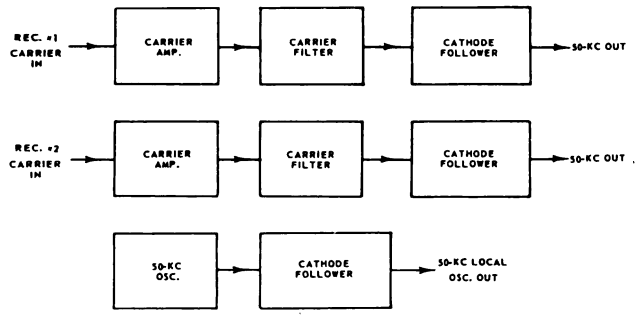

Figure 8-28. —Block diagram of r-f amplifier.

Signal comparator (1) amplifies the 50-kc carriers from two receivers operating in diversity and selects the stronger of the two signals (see fig. 8-29). Circuits are also provided for supplying the AGC voltage for the amplifier-converter units and the i-f amplifier units, plus a slow AGC voltage is used within the comparator for its own amplifier circuits. A carrier-operated-noise-silencer (CONS) voltage is also provided for the motor-tuned r-f oscillator.

An r-f gain control on the front panel adjusts the over-all gain by adjusting the AGC voltage. The rate (fast or slow) at which the diversity changeover occurs may be selected by means of a front-panel mounted comparing control. The AGC time constant is adjustable (off, fast, med, slow) from the front panel by means of the AGC switch.

A carrier level meter is mounted on the front panel to provide an indication of the strength of the received carrier. Two neon indicator lamps are used to indicate which of the two receivers is controlling the equipment.

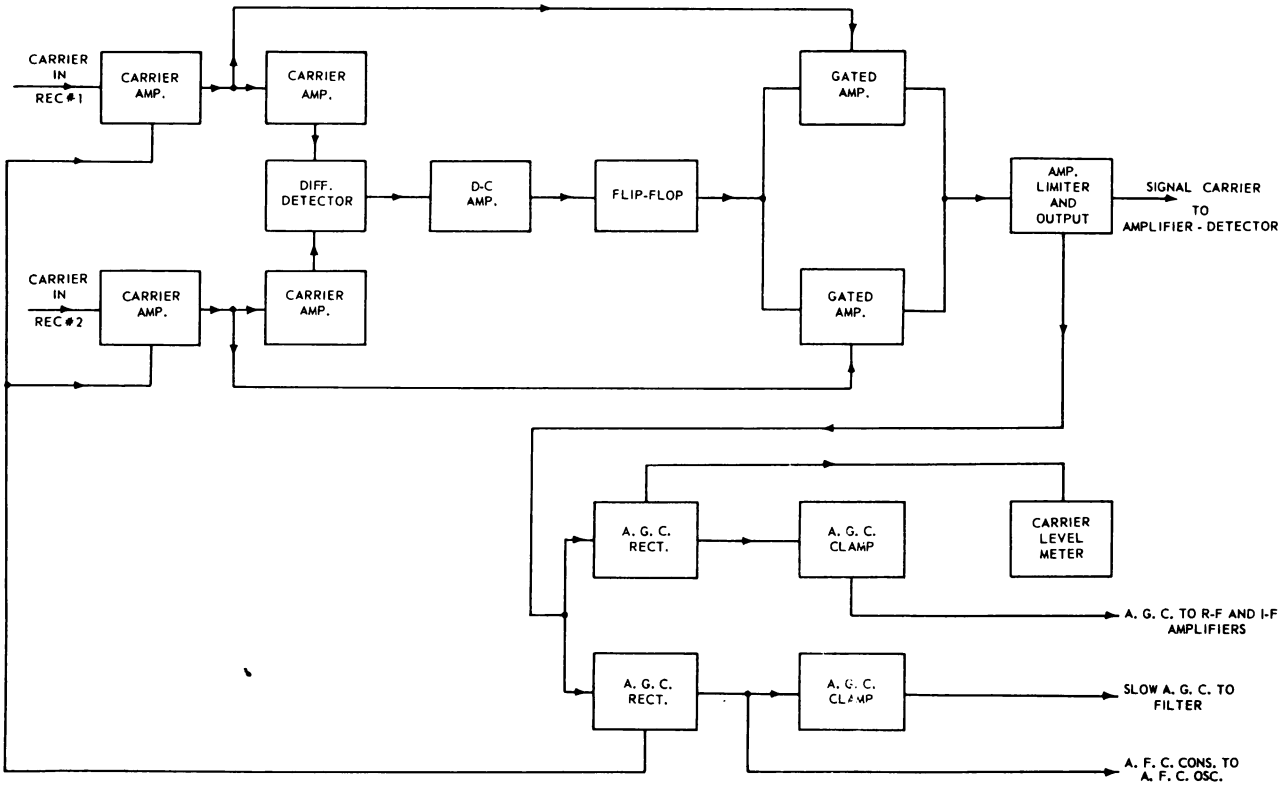

Figure 8-29. —Block diagram of signal comparator (1).

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QUIZ

1. In a conventional a-m system where both sidebands and the carrier are transmitted, what is the relative amount of power in the sidebands compared with that in the carrier for 100-percent modulation?
2. A single sideband transmitter rated at 50 watts will produce the same signal intelligence (under ideal conditions) as a conventional a-m transmitter of what rating?
3. If only one sideband is used, the SSB system requires what relative width of r-f spectrum compared with that of the a-m systems?
4. What is the permissible frequency variation for SSB systems compared with that of a-m systems?

Refer to figure 6-3 for questions 5 through 10.

5. What is the frequency of the audio modulating signal in the SSB transceiver?
6. What is the function of block 4?
7. What is the function of block 6?
8. What is the function of block 17?
9. At what point in the transmitter section may the carrier frequency be inserted if desired?
10. The combined signal (carrier and lower sideband) can be detected by what kind of receiver?
11. What is the relation between the nominal carrier frequency and the frequency of the channel-frequency crystals (fig. 8-4)?
12. What is the function of the tuning capacitor in the V-18 stage (8-4)?
13. What action maintains point B (fig. 8-6) at ground potential during most of the half cycle when E_{a-g} decreases from 148 volts to 50 volts?
14. What are two common methods of generating single sideband signals?
15. In the balanced modulator (fig. 8-7), what is the phase of the 250-kc signal on the V1 grid with respect to that on the V2 grid?
16. On what does the amount of carrier suppression depend (fig. 8-7)?
17. In figure 8-8, A, the upper sideband from balanced modulator 1 has what phase relation with respect to the upper sideband from balanced modulator 2?
18. In figure 8-8, A, which two components combine in phase in the add network?
19. Why are expensive narrow band filters not necessary in the phase shift system of SSB generation?
20. What is the range of the Q of narrow band crystal filters used in the filter method of SSB generation?
21. In the torsional-type mechanical filter (fig. 8-10), what is the effect of temperature on the resonant frequency?
22. The crystal lattice filter (fig. 8-12) is, in effect, what type of circuit?
23. In figure 8-14, what is the band of frequencies the audio system is designed to pass?
24. What type of signal is developed by the transmitting portion of the AN/FRC-10?
25. How is additional privacy obtained in the TEF transmitter (fig. 8-17) when double-sideband transmission is used?
26. What unit in the AN/FRR-10 (fig. 8-19) amplifies and separates the carrier from two separate composite signals and also supplies the crystal controlled, 50-kc reference for AFC control and demodulation?
27. Why does no sideband frequency inversion occur in the second converter (fig. 8-19)?
28. What is the function of the filters in the i-f amplifier (fig. 8-22)?
29. Which unit of the AN/FRR-10 (fig. 8-19) is used to amplify the upper and lower sidebands from two receivers operating in diversity reception and to select the stronger of the two signals?
30. In the amplifier detector (fig. 8-25), what type of detection is used to reinject the 50-kc carrier?
31. What is the function of the carrier-operated, noise-silencer amplifier (fig. 8-27)?
32. What unit in the AN/FRR-10 (fig. 8-19) amplifies and selects the stronger of the two 50-kc carriers from the r-f amplifier and supplies two types of AGC voltage and a carrier-operated-noise-silencer to other units?

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CHAPTER 9

BASIC PRINCIPLES OF COMMUNICATIONS MULTIPLEXING

INTRODUCTION

The number of communications networks in operation per unit of time throughout any given area is constantly increasing. In the past, each network has been required to operate on a different radio frequency. As a result, all areas of the r-f spectrum have become highly congested.

To a great extent, the maximum permissible number of intelligible transmissions taking place in the radio spectrum per unit of time is being increased through the use of multiplexing. Multiplexing involves the transmission of several intelligible signals during the same period of time normally required for the transmission of a single signal. Either of two methods of multiplexing may be used. These are time-division and frequency-division multiplexing, respectively.

GENERAL DESCRIPTION

TIME-DIVISION MULTIPLEXING

With a-m voice and tone communications, it is desired to transmit and receive the full 360° of each sine wave (fig. 9-1, A). However, an audio signal may be transmitted and received satisfactorily by periodically sampling the signal. The result of the sampling process yields a received signal such as that shown in figure 9-1, B. Slightly more than two samples per cycle of audio will give reasonably satisfactory results. In practical systems, 2.4 samples per cycle are usually taken. This concept of sampling forms the basis for time-division multiplex operation.

Figure 9-2, A, illustrates in a highly simplified form the basic principle of time-division multiplexing. Assume that a 3,000-cycle tone is applied to each of the six channels in the transmitter. Assume also that the rotating switch turns fast enough to sample, in turn, each of the

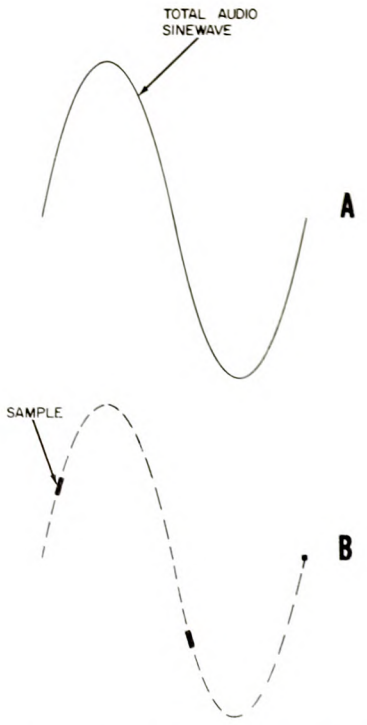

Figure 9-1. —Components of a sinewave.

six channels 2.4 times during each cycle of the 3,000-cycle tone. The speed of rotation of the switch must then be 2.4 x 3,000, or 7,200 rotations per second.

If the transmitter and receiver switches are synchronized, the signals will be fed in the proper sequence to the receiver channels. The transmitted samples from transmitter channel 1 will be fed to receiver channel 1. Thus, in the time-division method of multiplexing, many channels of audio are combined (with time spacing between components of the separate channels) to form a single output (multiplexed) chain. The chain is transmitted (via wire or radio facilities) to distant demultiplexing receivers, each of which functions to select only

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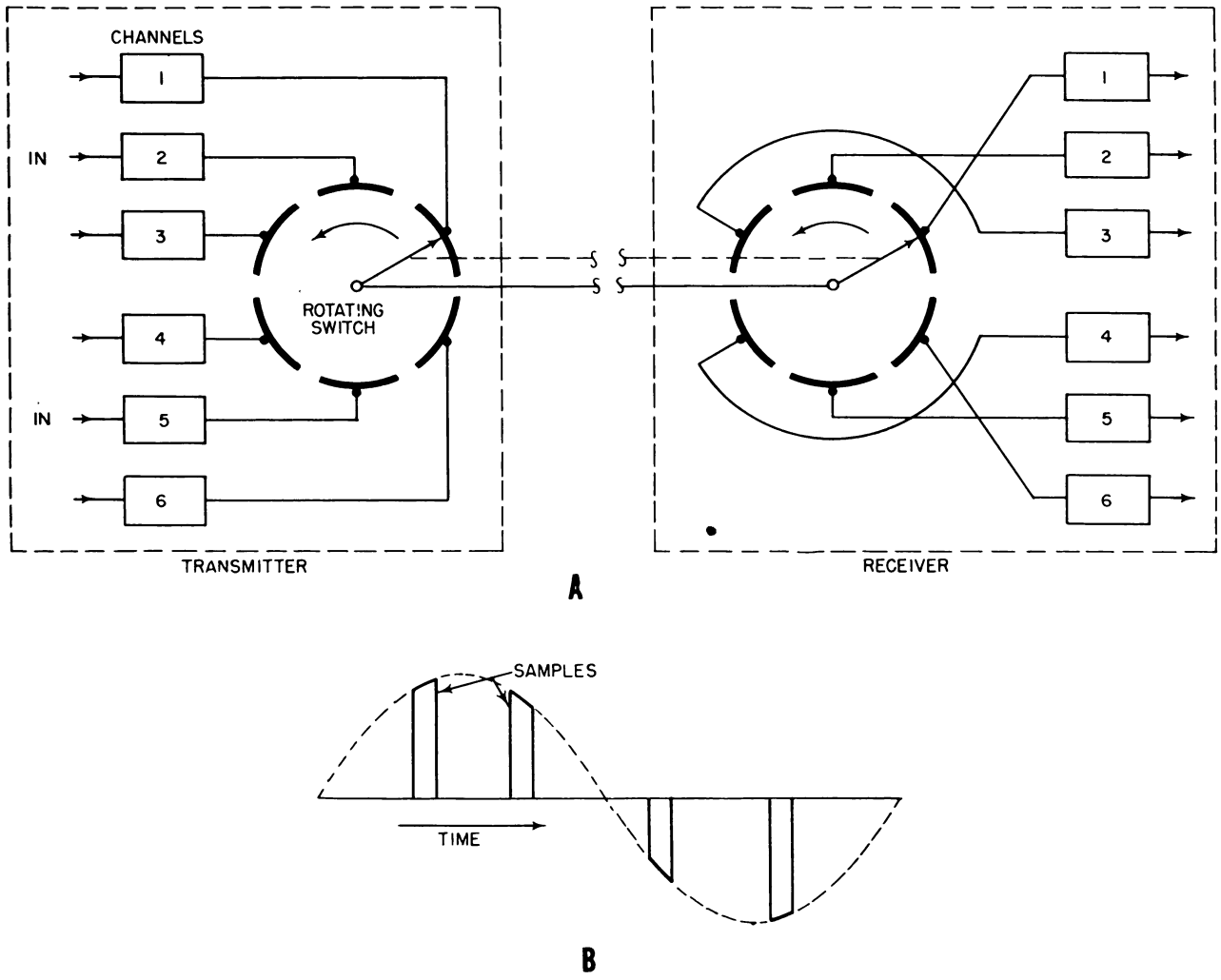

Figure 9-2. –Basic principle of time-division multiplexing.

the information pertaining to its particular channel.

A mechanical system is shown here for the sake of simplicity. In actual operation, any mechanical switching arrangement would not be able to provide the high speed of switching that is required for a multiplexing system. For this reason, electronic switching is used.

A sine wave sampled four times for each channel is illustrated in figure 9-2, B. In an actual transmission, segments from the waveforms in all of the channels will be interspaced with these four segments on a time-sharing basis. In other words, only one segment can be transmitted at a time; and the segments are taken in sequence from the waveforms existing in the six channels.

More than six channels (perhaps 24 or more) may be used. However, as the number of channels is increased, the width of each segment must be proportionately reduced. The great disadvantage in reducing the width of the pulse is that the bandwidth necessary for transmission is greatly increased. Decreasing the pulse width will decrease the minimum required rise time and increase the required bandwidth because of the increase in the number of harmonics contained in the sharper leading edge of the pulse.

The bandwidth is also affected by the shape of the sampling pulse and the method of varying the pulse to carry the modulation. The methods used include pulse amplitude modulation, pulse width modulation, and pulse position modulation. Pulse amplitude modulation is the type illustrated

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in figure 9-2. These methods are treated in more detail later.

In time-division multiplexing, to use time in the most efficient manner, the bandwidth necessary to transmit n channels would be only slightly wider than n times that necessary for one channel. However, the minimum bandwidth is not generally utilized.

FREQUENCY-DIVISION MULTIPLEXING

Frequency-division multiplexing (fig. 9-3) is the older of the two methods of multiplexing. In this system, different subcarrier frequencies are modulated by the signals of different channels, transmitted over the same cable (in the case of cable transmission) or on the same radio frequency carrier (in the case of radio transmission), then separated by filters before being demodulated. The total bandwidth required for a frequency-division multiplexing system is the sum of the bandwidths of the individual channels, plus the sum of the necessary guard-band frequencies between channels.

The resultant signal in each channel will be approximately the same as that which would be produced by the use of a single independent channel, except that nonlinear elements in the system may cause mixing of the various carrier frequencies and crosstalk between channels. For cable transmission, frequency-division multiplexing, is a satisfactory method because equipment is available which gives good fidelity, good signal-to-noise ratio, nearly constant gain, and very low nonlinear distortion.

Frequency-division multiplexing is not as satisfactory for radio communication because (1) the signal-to-noise ratio of a radio channel is lower than that of a wire channel, (2) the gain may vary considerably with atmospheric conditions, and (3) nonlinear distortion is difficult to eliminate.

A greatly simplified block diagram of a frequency-division multiplexing system is illustration in figure 9-3. For simplicity it may be assumed that the signals are transmitted over a cable; nevertheless a radio carrier could be used.

Channels 2 through 6 each employ an oscillator whose frequency is heterodyned with the audio input to the respective channels in a balanced modulator. The sum frequency produced as a result of the heterodyning represents a subcarrier where radio facilities are used. The filter used in each channel allows the passage of the upper sideband to modulate the transmitter r-f carrier. Channel 1 does not employ an oscillator, and therefore feeds its audio input (between 300 and 3,000 cycles) directly to the transmitter.

The bandwidth and position (with respect to frequency) of each of the six channels are indicated at the bottom of the figure. The total spectrum that may be occupied in this particular example is 19.5 kc.

Only the upper sideband (sum frequencies generated in the balanced modulator) is utilized in each channel. The suppressed carrier frequency is indicated by a dashed line between the various channels. This is the frequency used to generate the sideband indicated to the right-hand side of the line. For example, 3.3 kc is the oscillator frequency for channel 2. Although the upper sidebands are transmitted in this example, the lower sidebands could have been transmitted just as easily.

The channel 2 circuits at the transmitter and the receiver are used as examples in the block diagram of figure 9-4. For simplicity, a 1-kc tone is assumed to be the channel input.

At the transmitter, the 1-kc signal is fed to the balanced modulator; a 3.3-kc signal is also fed to the balanced modulator from the channel 2 oscillator. The balanced modulator supresses the carrier and generates upper (4.3 kc) and lower (2.3 kc) sidebands. Only the upper sideband is passed by the channel 2 filter.

The channel 2 demodulator in the receiver accepts the 4.3-kc signal and heterodynes this input with the channel 2 oscillator signal (3.3 kc). The carrier is suppressed, and the two sidebands (7.6 kc and 1 kc) are fed to the channel 2 demodulator filter. The filter passes the 1-kc signal to the reproducer.

TIME-DIVISION MULTIPLEX SYSTEMS

Either the amplitude or the frequency of the transmitted signals (samples) from the respective channels may be varied to affect modulation. Three practical methods of pulse modulation are used. These are pulse amplitude, pulse width (pulse duration), and pulse position modulation. In either case, a modulating pulse is made to vary at a rate dependent upon the frequency of the audio. Each method is described in detail later in this chapter.

The sampling process allows modulating pulses (2.4 or more sampled at a fixed rate) to be introduced to the r-f transmitter carrier, or

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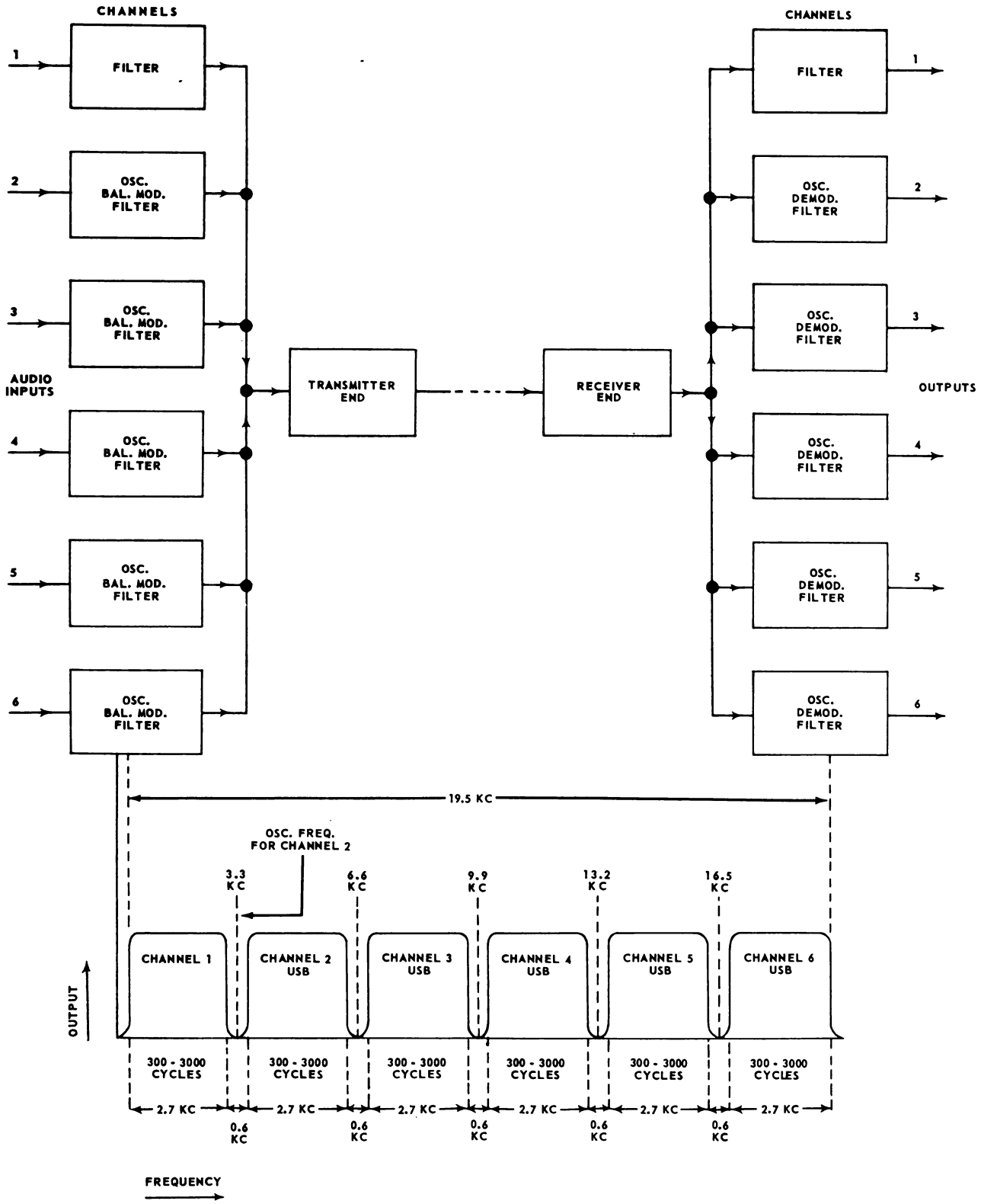

Figure 9-3. –Block diagram of a frequency-division multiplexing system.

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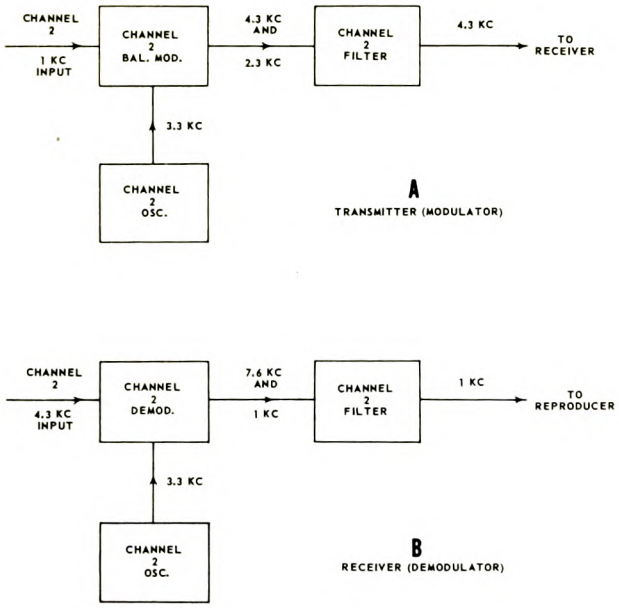

Figure 9-4. –Block diagram of a modulator and demodulator units.

to the wire cable connecting the transmitting and receiving equipments. At least 2.4 of these samples must be transmitted for each cycle of the audio in order to produce an intelligible signal at the reproducer.

If the rate of sampling throughout the multiplex system is adjusted to 9.3 kc, then the maximum allowable audio frequency = $\frac{9300}{2.4}$ = 3875 cps. This frequency represents the highest audio frequency that may be used. In practice, it is desirable to obtain a higher fidelity than can be realized with the minimum number of samples per second. For this reason, filters would be employed in this system to eliminate frequencies above about 3000 cps.

BLOCK DIAGRAM

Sampling the audio signal of several channels through the use of electronic switching can be accomplished in several ways. A block diagram analysis of one method used in time-division multiplex is shown in figure 9-5. In this system, it is first necessary to develop a voltage that varies linearly with time. The sawtooth (time base) generator is of the bootstrap type, which uses a linearity stage to improve both voltage output and voltage linearity. Its output varies sufficiently linearly with respect to time to establish the basis for time-division multiplex.

The sawtooth generator output voltage (through the rise time) is applied to the grid of a normally cutoff sawtooth selector stage. Conduction of this stage occurs when the sawtooth amplitude reaches a potential slightly less than the bias voltage on the selector. When the sawtooth voltage becomes sufficiently high to cause conduction of the selector, a trigger pulse is produced at the grid of a one-shot multivibrator. The multivibrator output is applied along dual paths to the sawtooth oscillator and discharge tubes to cause the sawtooth flyback. The bias on the selector may be manually varied to change the sawtooth amplitude as will be shown later.

The duration of the sawtooth waveform is adjusted to exactly 107 μs. The sawtooth selector stage allows a linear rise time from approximately +50 to +215 v (165 v corresponding to the total rise time).

The sawtooth wave form is fed along paralleled paths to the channel selectors, one of which is contained in each channel of the multiplex system. The individual selectors are biased each at a different level so that they are energized (at different times) by the sawtooth voltage rising at the respective grids.

As each channel selector is energized, it feeds a pulse to a channel modulator, causing this circuit to produce a channel pulse. The channel pulse thus produced (in this channel alone) is then modulated by the incoming audio in accordance with the type of pulse modulation used.

The multivibrator output trigger, which is fed to the sawtooth oscillator and discharge tubes, has a duration of 4.75 μs. Thus, these stages are allowed to conduct for this entire period, which accounts for the delay between the decay of one sawtooth waveform and the beginning rise time of the next.

Included in the transmitter output is a marker pulse (sometimes called a sync group). These pulses may be of any form but are usually of a nature that makes them easily distinguishable from the modulation pulses at the receiving terminal of the multiplex system. A circuit used to produce a marker pulse will be described presently. The marker pulse (or pulses), along with the total modulation pulses from each channel (in time) presented to the transmitter carrier during one complete cycle of the sawtooth waveform, constitutes a video frame.

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The negative gate from the multivibrator (occurring at the same time as the positive pulse that produces the sawtooth decay) gates a Hartley oscillator (marker pulse) circuit to cause the generation of the marker pulses. The sinusoidal oscillator output of the marker generator is acted upon to produce four .5-μs pulses spaced 0.8 μs apart. These pulses are passed through a cathode follower (impedance matching) circuit to the transmitter. Both marker pulses and channel pulses can be coupled along the same line to the transmitter, because each of the pulses is added to the video frame at a different time interval. The primary purpose of the marker pulses is to act as a synchronizing voltage between the receiving and transmitting multiplex terminals.

CIRCUIT ANALYSIS OF A TIME-DIVISION MULTIPLEX SYSTEM

Multiplex Waveform Generator

The boot-strap sawtooth oscillator (fig. 9-6) comprises tubes V1, V2, and V3. This circuit arrangement is used mainly because it supplies a very linear sawtooth output voltage with respect to time. In the conventional sawtooth oscillator circuit, the electrically linear portion of the output is generally equal to about ten percent of the voltage applied to the circuit. Thus, to produce a 165-v linear sawtooth rise, a B supply voltage value of 1650 v would be required. Although this value of voltage could be easily obtained through step-up power transformer action, the boot-strap circuit will produce the desired 165-v linear output with a 360-v B supply voltage applied. This represents a considerable decrease in required voltage over the conventional circuit.

To understand the boot-strap principle of operation, consider the action that takes place when the B supply voltage is applied. V1 is held beyond cutoff by the -105 v at its grid. C1 and C2 begin to charge to a positive value. This positive potential is fed to the control grid of the linearity tube, V2. V2 is a conventional cathode follower with a relatively high value of cathode resistance R5 (12 kilohms).

With the V2 grid more positive, a higher current passes through R5. As a result, a more positive voltage is developed at the cathode of V2, which is applied (as a transient voltage) through C3 to the R2-R3 junction. This rise in voltage is linear and is applied to C1 and C2 via R2. The boot-strap action can be seen at this point since it is the voltage developed across R5 that boosts the charge of the capacitors C1 and C2 to a higher value.

The voltage fed to the V2 grid rises linearly as a result of the boot-strap charging of the capacitors C1 and C2. Thus, the R5 voltage increases linearly with respect to time.

Another way of showing the linearity of the sawtooth waveform is to note that the rise in voltage at the V1 plate is the same as the rise in voltage at the R2-R3 junction. Thus, the current through R2 is constant during the charge time of C1-C2. Since charging current varies directly with its rate of change of capacitor voltage, if the current is constant, the rate of voltage rise is constant. This action is described in similar boot-strap circuits in subsequent chapters of this training course.

The sawtooth selector tube, V4, receives an adjustable bias potential at its cathode, obtained from a voltage divider (comprising R7, R8, and R9) connected from the B supply to ground. The grid receives the linearly rising sawtooth voltage from R5. When the amplitude of the sawtooth voltage reaches a value sufficient to overcome the V4 bias, the tube conducts, allowing its plate voltage to drop. The voltage change is differentiated by C4 and R10 to form a sharp negative peak, which cuts off the normally conducting inverter amplifier, V5. The resulting positive trigger, differentiated by C7 and R15 serves as the trigger pulse for the one-shot multivibrator (V-6) circuit. This signal is also fed through C6, and developed by R12 and R5 as a positive-going regenerative feedback at the V4 grid. This action hastens the saturation of V4, which, in turn, decreases the time to cut off V5. The latter action sharpens the trigger fed to the V4 grid.

The B side of the multivibrator tube, V6, is normally conducting as a result of a positive potential applied at its grid. This potential is obtained from a voltage divider comprising R22 and R23, which is connected between the B supply and ground. The voltage developed across the 1200-ohm cathode resistor, R20, as a result of the V6B conduction is sufficient to maintain the V6A grid potential well beyond the cutoff point.

Upon arrival of the positive trigger from V5, the V6A grid is raised above cutoff. Conduction of this tube section causes a decrease in the V6A plate voltage and a rise in the voltage across the common cathode resistor, R20. The

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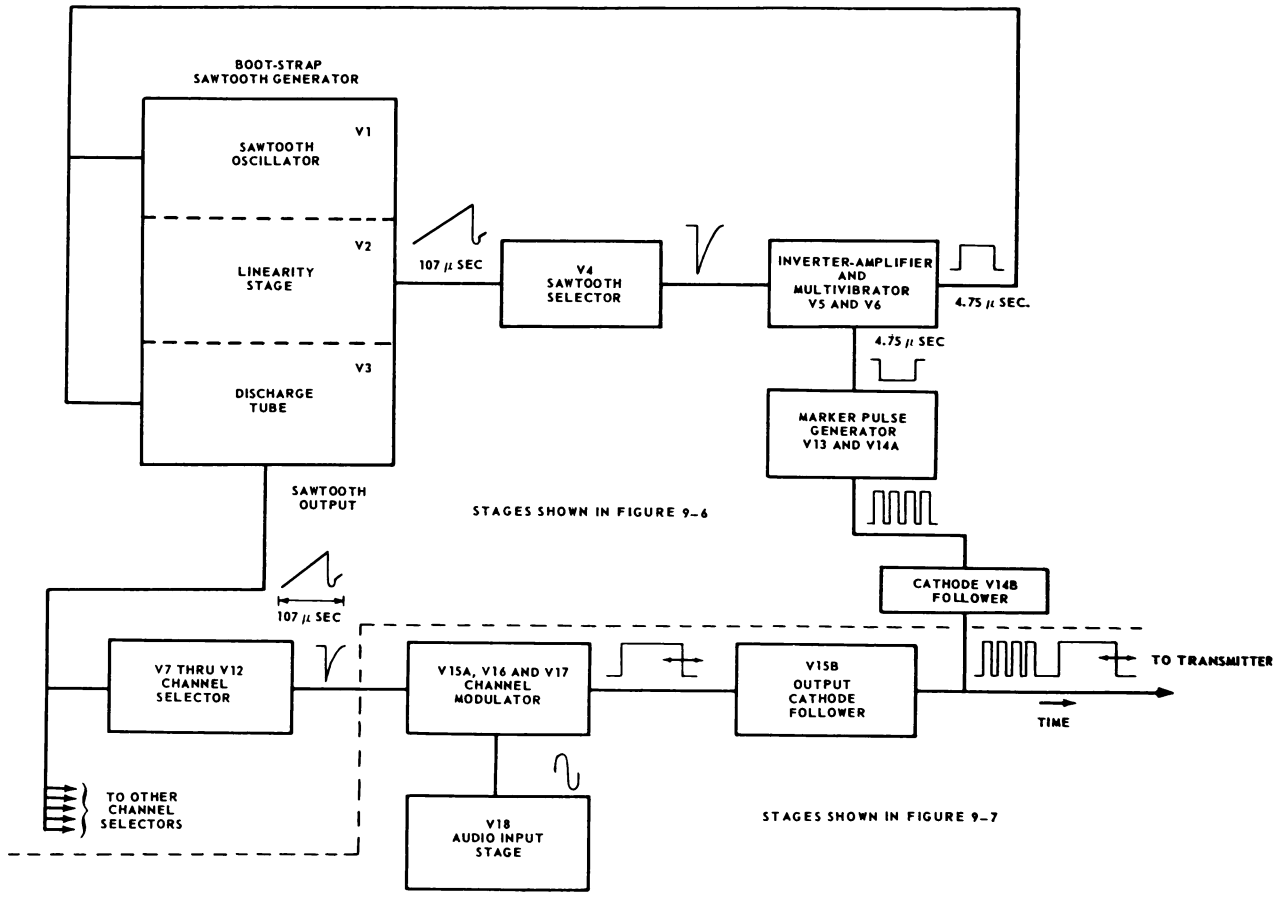

Figure 9-5. –Time-division multiplex block diagram.

drop in the V6A plate voltage is fed through C9 to initiate the cutoff of V6B. The action is cumulative until V6A conducts saturation current and V6B is at cutoff.

As a result of the V6B plate voltage rise, a positive 4.75-μs pulse is developed at the grid of the sawtooth oscillator, V1, and the discharge stage, V4, by R1 and R6, respectively. This action simultaneously triggers both tubes into conduction. With V1 conducting, C1 and C2 discharge through the V1 conducting resistance.

The positive trigger at V3 allows the tube to conduct to effectively short R5. Thus the duration of the decay is equal to the time required for V3 to reach saturation. Since the positive potential at the V3 grid is continuous for approximately 4.75 μs, the sawtooth waveform amplitude will remain at about +50 volts (with V3 conducting) throughout this period. At the end of the pulse from the multivibrator, V1 and V3 are again cut off by the high bias voltage.

The sawtooth output of the multiplex waveform generator is applied simultaneously to the control grids of the channel modulators, V7 through V12. Each channel is biased at a different d-c level, all beyond cutoff of the tubes, and at a point that lies within the linear portion of the sawtooth rise. Thus, each channel selector conducts at a different time. The bias voltage applied to the respective cathodes is obtained from a voltage divider (comprising R24 through R30) connected between the B supply and ground.

Because of the arrangement just described, each channel selector is caused to produce a trigger pulse for its particular channel at different times along the 107-μs sawtooth sweep voltage rise. The resulting drop in plate voltage of the channel selector initiates the development of the channel pulse. The channel pulse is modulated and allowed to enter the video frame through a process described later.

Once rendered conductive, each channel will continue to conduct throughout the remaining

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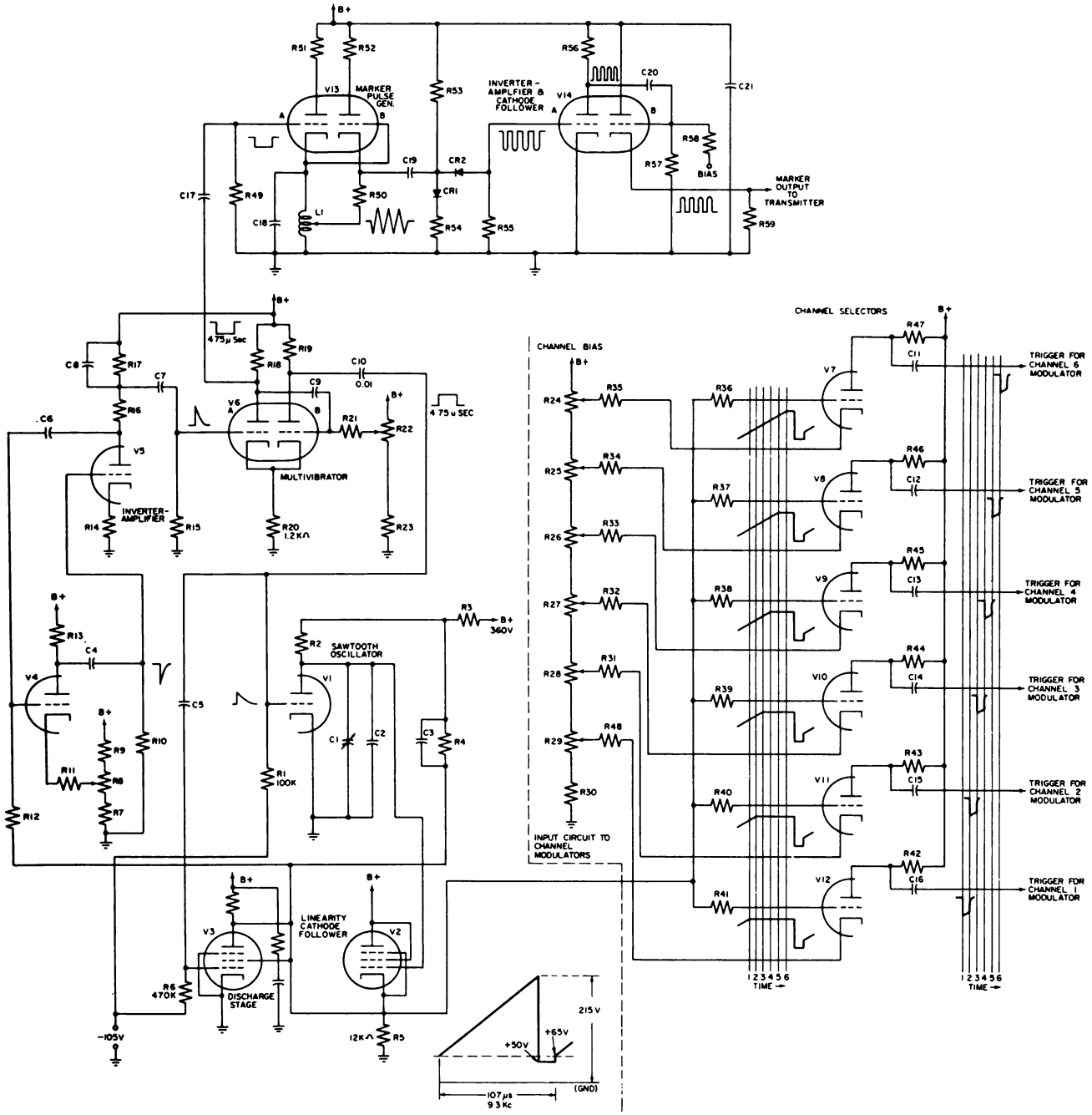

Figure 9-6. –Time-division multiplex operation.

portion of the sawtooth rise. During the sawtooth decay, all selector tubes cease to conduct, and are held nonconductive (by the bias at their respective cathodes) until they are again made conductive (in time sequence) by the following sawtooth waveshape. Thus, the sawtooth waveshape serves as the master (or controlling) signal for the multiplex system.

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Marker Pulse Generator

It may now be recalled that the multivibrator, V6, supplied a positive 4.75-pulse, which initiated the sawtooth decay. The rise time of the following sawtooth was delayed throughout the 4.75-μs period. The negative portion of the V6 output (from V6A) is applied through C17 to the control grid of the marker pulse generator, V13. As a result of zero bias, V13A normally conducts saturation current through coil, L1. C18 and L1 make up the tuned (780-kc) tank circuit for the pulsed Hartley oscillator stage, V13B.

When no pulse is applied at V13A, saturation current constantly passes through L1 to prevent oscillations. The negative trigger cuts off V13A, allowing the field about L1 to collapse, and oscillations begin. The less positive (negative-going) potential now developed at the V13B grid is regenerative and thereby aids the oscillatory action.

When the field about L1 completely collapses, C18 begins to discharge. The rising current through L1 causes a positive potential to be developed at the V13B grid, which aids the oscillations. The duration of one complete cycle of the 780-kc oscillation is approximately 1.3 μs. Each half cycle duration is then approximately .65 μs. Thus, during a 4.75-μs period (the duration of the trigger pulse) slightly more than three and one-half cycles can be completed. At the end of the negative trigger pulse from V6A, V13B again conducts saturation current and the V6B oscillation ceases.

The oscillator output during the trigger pulse is developed across R50 and applied through C19 to the CR1-CR2 junction. CR1 is normally conducting as a result of a connection to the B supply through voltage divider R53, CR1, and R54. The potential developed at the CR1-CR2 junction is approximately +2 v.

When the pulse from the oscillator exceeds 2 v in the negative direction, the positive potential at the CR2 cathode is sufficiently reduced to allow the CR2 conduction. Thus, the portion of the total V13 output, which is greater than -2 v, is developed across R55 and applied as a changing negative-going potential to the grid of the inverter-amplifier, V14.

The duration of the individual pulses applied at the V14 grid as a result of the biased clipping is reduced to .5 μs. The spacing between pulses is increased to approximately 0.8 μs.

V14A amplifies and inverts the pulses and applies them through C20 to the cathode follower (V14B) control grid. The marker output is developed across R59 and applied to the transmitter during the time that the voltage across R5 is held constant on 50 v.

Channel Modulator

Each channel selector output (fig. 9-6) is fed to its associated channel modulator (fig. 9-7). The channel selector output is received at the control grid of the inverter-cathode follower, V15, which serves as the input stage of the associated channel modulator. V15A is cathode biased and is therefore normally conducting.

When V15A cuts off, the plate voltage begins to rise, and parallel capacitors C24 and C25 begin to charge toward the B supply value through R69, R62, and the resistance of the B supply. In the quiescent condition (no modulation) the charge time is adjusted to 3.2 μs by trimmer capacitor C24.

As a result of a connection to the B supply, V16A normally conducts through R68, the grid to cathode resistance of V16A, and R65. The drop across R68 biases V16B beyond cutoff. When V16B conducts, in response to the positive trigger from V15A, the drop in plate voltage is applied through C29 to the V16A grid. This action drives V16A to cut off, and the corresponding plate voltage rise represents the leading edge of the channel pulse.

The V16A plate voltage rise causes C27 and C30 to charge toward the B supply value through R70, R66, and the resistance of the B supply. The voltage at the V17A grid rises exponentially as a result of the integrating action of C30. The V17B grid is connected to a -17-volt source. The negative voltage at the grid is taken from the R81-R74 junction. The R74 to ground potential is very small, allowing the positive voltage at the arm of R76 to effectively control the tube. R76 forms a portion of a voltage divider comprising R76 and R77, which is connected from the B supply to ground.

Because of the positive V17B grid voltage, a high current passes through the common cathode resistor, R73. This resistance is relatively large (considering the normal size of cathode bias resistors) so that a large cathode to ground voltage exists at both cathodes. The grid to cathode (bias) voltage of V17A (during quiesence) is large enough to keep the tube cut off. Note that the amount of voltage developed beyond the cutoff point of V17A is determined by the setting of the arm of R76, or by any potential, which is received at the V17B grid.

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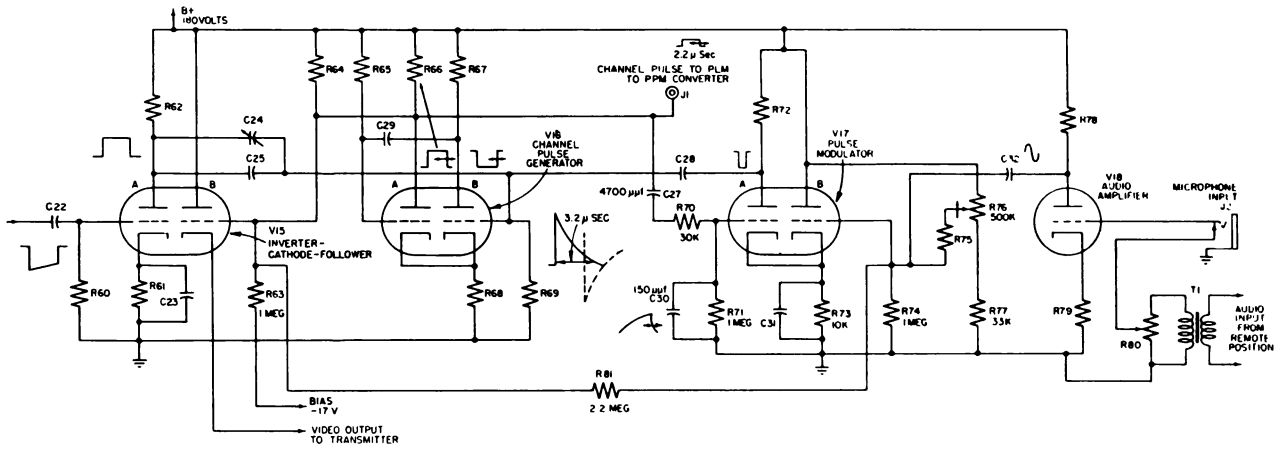

Figure 9-7. –Channel modulator, schematic diagram.

It is well at this time to review the train of events, which has taken place since the trigger from the channel selector was received. The negative-going channel selector input (leading edge of the pulse) drove V15A to cut off. C24 and C25 began to charge toward the B supply value through R69. The positive pulse developed at the V16B grid triggers the tube to conduct. The normal conducting time of V16B is 3.2 μs. The negative-going potential at the V16B plate is coupled through C29 to the V16A grid. The positive leading edge of the V16A (plate) output represents the leading edge of the channel pulse.

V17B normally conducts to hold V17A beyond cutoff. The charge of C30 (caused by the V16A plate voltage rise) moves positively (exponentially) at the V17A grid to eventually overcome the R73 voltage, and V17A conducts. The point (or time) at which V17A conducts is dependent upon the R73 potential, which, in turn, depends upon the setting of the arm of R76. The R76 arm is adjusted so that a 2.2-μs charge time of C30 is necessary to overcome the V17A bias.

The V17A conduction is accomplished by a drop in plate potential 2.2 μs after the leading edge of the channel pulse was produced. C28 discharges through R69 at this time to cut off V16B. The plate voltage rise of V16B charges C29 through R65, and V16A again conducts to form the trailing edge of the channel pulse.

Recall that the positive pulse at the V16B grid would normally allow conduction of V16B for 3.2 μs (as determined by the setting of C24) in the absence of the return negative trigger from V17A. However because the negative trigger is returned to the V16B grid, the width of the channel pulse is reduced to 2.2 μs.

If the potential at the arm of R76 is increased (clockwise rotation) the charge time of C30 necessary to overcome the increased V17A bias (across R73) will also be increased. Likewise, the time required before the negative trigger is returned from V17A to cut off V16B is also increased. Thus, a wider channel pulse is produced.

Conversely, if a less positive potential is applied from the arm of R76 (counterclockwise rotation) a shorter charge time of C30 is necessary to overcome the V17A grid-cathode bias. The negative trigger from V17A to cut off V16B occurs earlier, and the channel pulse width is reduced.

To modulate the channel pulse, audio is applied locally from a microphone jack, J2, or remotely by way of an audio transformer, T1. V18 is a conventional audio amplifier with degenerative cathode bias. The output of V18 is coupled through C32 and superimposed on the V17B grid voltage. The voltage alternately aids and opposes the potential at the arm of R76, which, in turn, increases and decreases the V17 voltage across R73.

The amplitude of the audio signal at the time of the channel pulse determines the magnitude of the V17A bias voltage, which must be overcome by the charge of C30 to allow V17A to conduct. If the audio is in its positive alternation, a larger voltage is developed by R73, and C30 must charge for a longer period. The negative trigger from V17A occurs later in time. The

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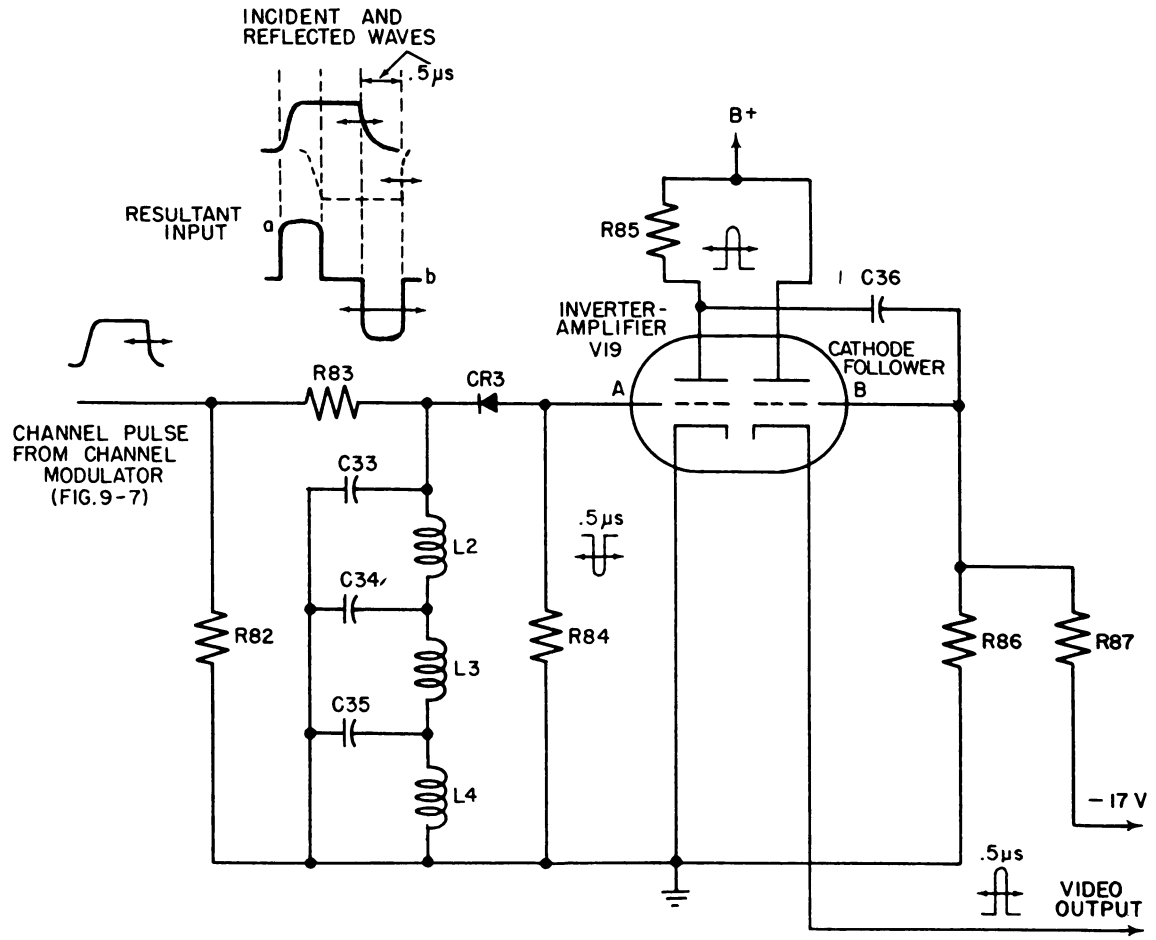

Figure 9-8. -PLM to PPM converter, schematic diagram.

width of the channel pulse is therefore greater than 2.2 μs.

Conversely, if the audio signal is passing through its negative alternation, at the time the channel pulse is produced, less bias voltage is developed by R73, and C30 charges for a shorter period. The channel pulse in this case will have a duration less than 2.2 μs, since the negative trigger from V17A occurs earlier than normal. Thus, it may be seen that the channel pulse produced is pulse width (or pulse duration) modulated.

The channel pulse, taken from the V16A plate, might be applied (through the proper impedance matching networks) to the transmitter. However, the width of the pulse is much greater than necessary to affect modulation. Since the intelligence is carried in only a small portion of the 2.2-μs pulse, considerable power will be lost during modulation. For this reason, the width of the pulse is reduced and the method of modulation altered as will be described later.

Pulse Length Modulation to Pulse Position Modulation Converter

The pulse length modulated signal from the channel modulator is applied at the input of the pulse length modulated (PLM) to pulse position modulated (PPM) converter (fig. 9-8). The circuit functions to reduce the width of the output modulated signals, and to change the pulse length modulation to pulse position modulation. This action is not indicated in the block diagram of figure 9-5.

The trailing edges of the input pulses change position relative to the percent of modulation of the pulse. One-hundred percent modulation cor-

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responds to ± 1μs variation in the pulse length.

The input pulse is applied through R83 to a 0.5-μs - delay line comprising inductors, L2, L3, and L4, and capacitors C33, C34, and C35. The lower end of the line is shorted to cause a reflected wave (0.5 μs delayed) 180° out of phase with the input. Thus, when the signal is applied to the line, 0.5 μs is required before a reflected pulse (of equal amplitude but opposite direction) is sent back through the line to oppose the input signal. During this time the portion of the pulse at "a" is produced. At the end of the incident wave, no additional energy is fed down the line and the resultant output is an unopposed signal 180° out of phase with the input, having a duration of 0.5 μs.

If the signal input is unmodulated, the position of the leading edge of the negative pulse will always be 2.2 μs from the leading edge of the input pulse. If the incident wave is caused to cease after 1.8 μs (as may be the case during modulation) the negative portion of the pulse at "b" will be produced by the delay line earlier in time. Conversely, if through modulation the channel pulse length is increased to 2.6 μs the pulse at "b" will be produced from the line later in time. In this manner, the position of the pulse at "b" is made to correspond with the trailing edge of the pulse length modulated signal input.

CR3 conducts to pass the negative pulse position modulated portion of the delay line output across R84. The positive pulse at "a" is not modulated since its position (in time in the video frame) is always the same. CR3 blocks the passage of this portion of the delay line output.

Although the pulse is represented as having a left-right motion, during modulation, only one pulse is produced for each modulator during one sawtooth sweep from the waveform generator (fig. 9-6). The sawtooth frequency should be recalled as being 9.3 kc. A pulse recurring at this rate will appear (during modulation) as a constantly moving pulse passing through all positions from maximum left to maximum right and so on.

Signals developed across R84 are applied to the grid of the inverter-amplifier, V19A. The negative inputs cause a V19A plate voltage rise, which is applied through C36 to the cathode follower (V19B) control grid. One cathode load resistor (common to all channels) is located in the transmitter. This resistor (not shown) functions to introduce the marker pulses (discussed earlier) and the separate modulated channel pulses (in time) to the transmitter carrier modulating circuits.

Multiplex signals may be transmitted by the same methods employed for sinusoidal amplitude modulated signals. However, one modulated envelope will have a duration equal to that of the master oscillator (sawtooth) frequency (fig. 9-9) of 107 μs, and will contain a sample from each channel of the multiplex system.

If normal modulation methods were used, unnecessary power would be lost during intervals when no pulse was being transmitted. Therefore, to avoid this loss the separate output pulses are made to key the transmitter on and off to produce a pulsed carrier output.

DEMODULATION IN A TIME-DIVISION MULTIPLEX SYSTEM

Block Diagram

The signal from the multiplex receiver is applied to the demodulator chassis via input jack, J1 (fig. 9-10). This input is amplified and shaped in stages V1 through V3. A peaking at the output of V3 insures a 0.5-μs duration for each pulse in the video frame.

V4 acts as an amplifier and cathode follower to supply two output signals. The plate output is negative-going, and is applied to an integrating circuit. The cathode output is a positive video frame, which is coupled simultaneously to each of the channel demodulators.

The time constant of the integrating circuit between V4 and V5 is such that the sync pulse is integrated into one pulse whose total duration is approximately 4.35 μs. The individual channel pulses are reduced in width by this circuit. V5 amplifies this input and applies it to another integrating circuit in which only the relatively long duration of the sync pulse is sufficient to charge the capacitor in the integrating circuit high enough to cause the V6A conduction. Thus, one output pulse is produced by V6B, which corresponds in time to the sync pulse in the video frame.

It should be recalled that the marker (or sync) pulses just described are produced by the multiplex system at a time that corresponds to the rest (or delayed) time of the multiplex sawtooth waveform. The beginning of the following sawtooth takes place immediately after the end of the generation of the marker group.

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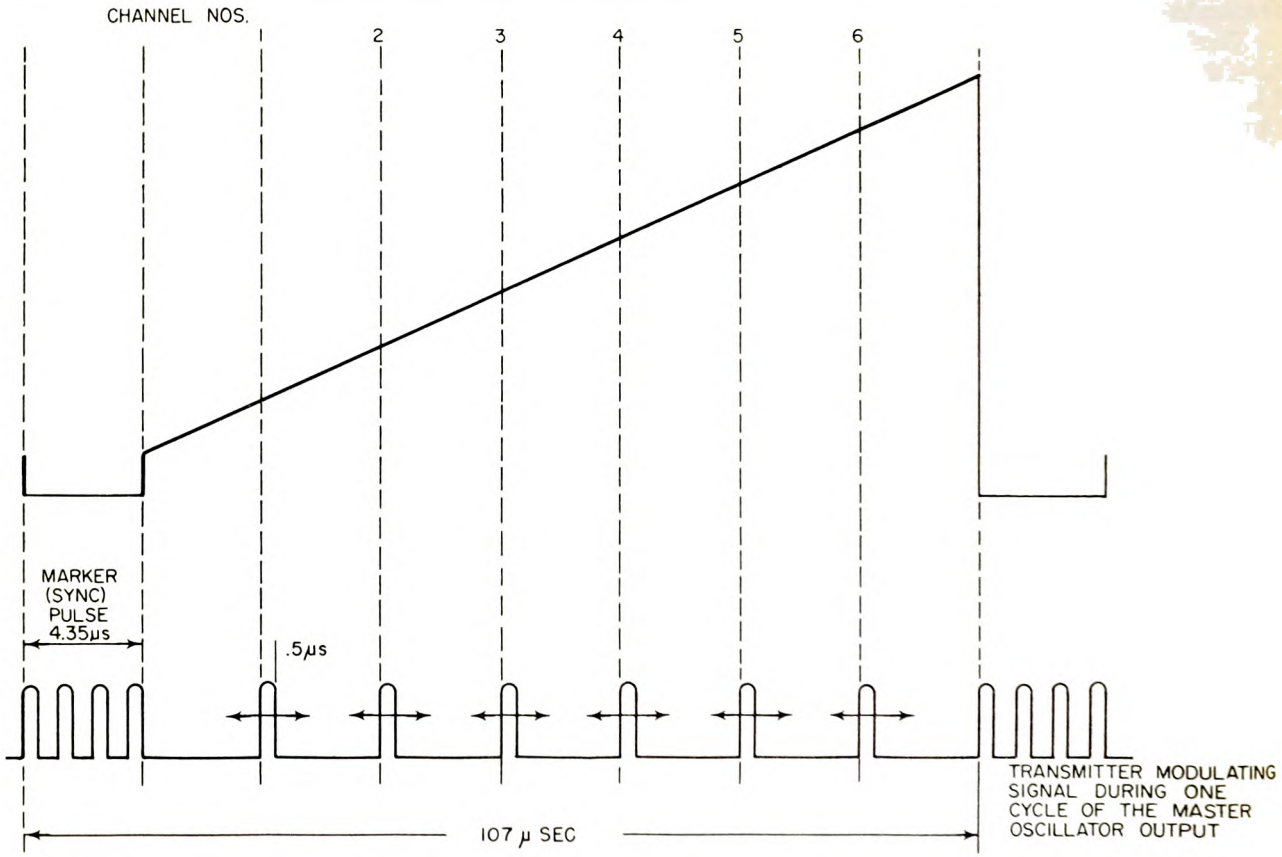

Figure 9-9. –Relationship of master oscillator output to composite video frame.

The position of the trigger pulse produced by V6B is made to correspond with the trailing edge of the marker pulse trigger (see V13, figure 9-6). The output of V6B is therefore made to trigger the beginning of the sawtooth in the demultiplex system, and the sawtooth rise of both the multiplex and demultiplex systems are caused to occur at the same time.

The sawtooth generator circuits comprise tubes V7, V8, and V9. These tubes are connected in a boot-strap arrangement, which is very similar to that already discussed in the multiplex system (fig. 9-6). The sawtooth output is made available at J2 for test purposes. J3 represents the output jack from which the sawtooth output is connected to each of the channel demodulators (in parallel). This output triggers the demodulator channels in synchronism with the time of arrival of the individual channel pulses in the video frame.

The channel demodulator, V10 (fig. 9-11) receives the sawtooth waveform from the demultiplex waveform generator. This stage is biased to conduct at a specific point along the rise of the sawtooth at the grid. When V10 conducts, a negative output is applied to a trigger stage. V12A amplifies and inverts the grid input, which is used to trigger a blocking oscillator, V12B. The blocking oscillator output is shaped and applied to the gating-out amplifier, V13.

V13 functions as a coincidence stage, which requires the simultaneous application of the channel pulse, and the blocking oscillator input. In this stage, the pulse position modulation is changed to pulse amplitude modulation by affecting the amplitude of the V13 plate output.

Stage V14 removes the modulation component, and applies its output to a 9.3-kc suppression circuit, which removes the high frequency component of the output.

The audio amplifier, V15, increases the amplitude of the audio, and applies its output to audio output jack J3. The output audio has a bandpass of approximately 3 kc.

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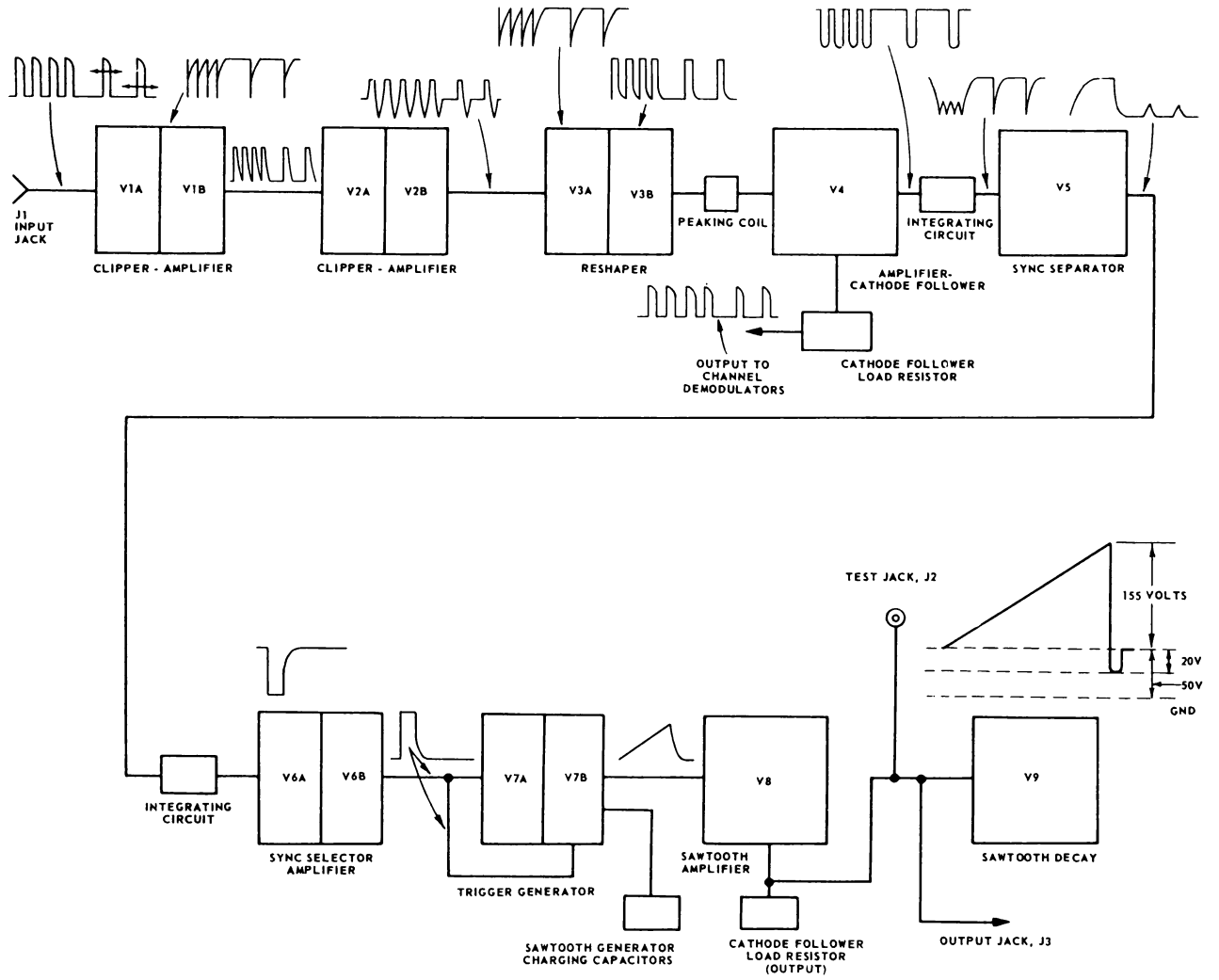

Figure 9-10. –Demultiplex waveform generator, block diagram.

Demultiplex Waveform Generator

A conventional superheterodyne receiver (not shown) amplifies and detects the multiplexed signals and passes them to the demultiplexer circuits. The successful reproduction of the signals in the demultiplexer depends primarily upon the degree of synchronization, which can be maintained between the multiplex and demultiplex equipments. The marker pulse generated by V13 in the multiplex waveform generator (fig. 9-6) always occurs at a constant rate (9.3 kc) and in a fixed position in time. The pulses represent the beginning of the video frame, and are utilized by the demultiplex system for synchronization purposes.

Input signals to the demultiplex waveform generator (fig. 9-12) are amplified in V1A and applied through C2 to the V1B grid. These stages act as amplifiers for weak input signals, and may be overdriven by strong signals to become limiters.

The positive output pulses of V1B are applied to the V2A grid and are received at V2B as negative input signals. The positive output signals from V2B are applied through C7 to shock L1 into oscillations with its distributed capacitance. L1 reshapes the separate pulses during positive portions of the signals, but is severely damped

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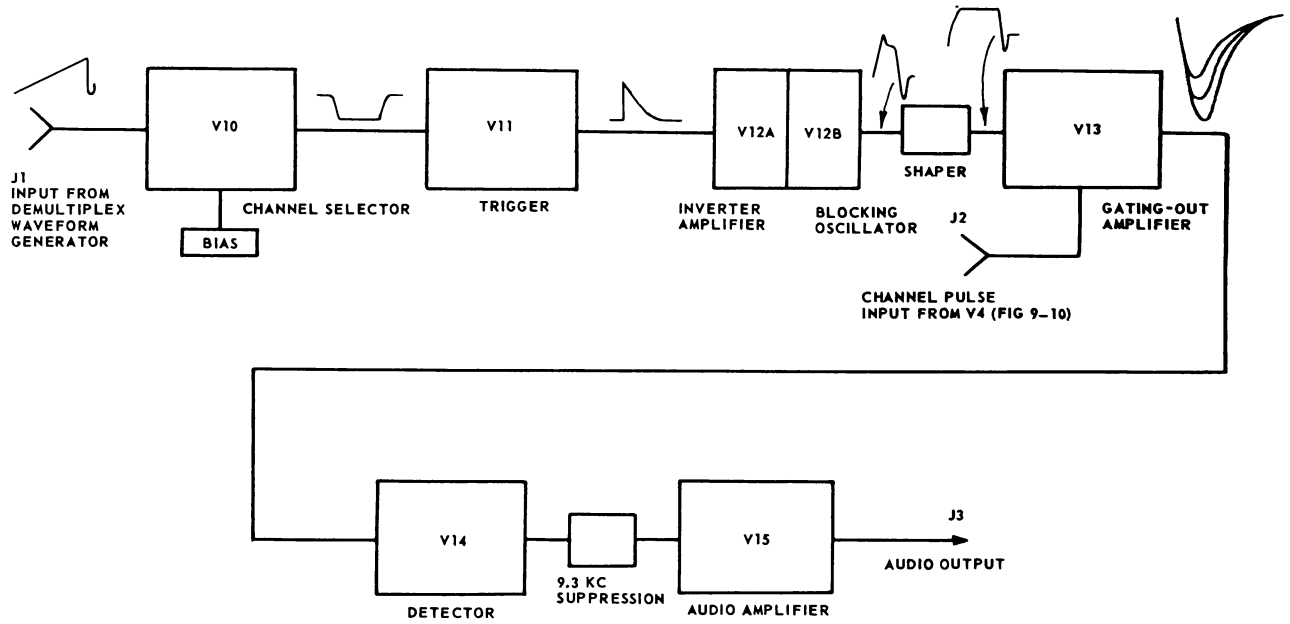

Figure 9-11. –Channel demodulator, block diagram.

during negative alternations by CR1. The diode conducts to short negative signals around L1.

V3A amplifies and inverts the signals. The plate of V3A and the control grid of V3B are directly connected, allowing V3B to conduct. Negative signals from V3A are integrated across the grid to ground distributed capacitance (Cd) to produce a positive sawtooth at the V3B plate. L2 acts as a pealing coil in the V3B plate to produce a pulse width of 0.5 μs. These pulses are applied through C10 to the grid of the amplifier-cathode follower, V4.

The plate resistor, R20, of V4 has twice the ohmic value of the cathode resistor, R22. The cathode output is applied via J2 to the demodulator circuits. This output represents an amplified video frame which is to be demodulated by a process described later.

The plate output of V4 is used to trigger the remaining circuits of the waveform generator to produce a demultiplexing sawtooth for the demodulators. This output is very similar to the sawtooth waveform of the multiplex system, (fig. 9-9) and is used to trigger the demodulator channels at specifically timed intervals, which correspond (in time) to the position of a particular channel in the multiplex video frame.

The V4 plate signals (fig. 9-12) are coupled by C12 to the CR2 diode, which is a portion of a voltage divider connected from the B supply to ground, and comprising R23, CR2, and R24. The amount of the CR2 conduction (and therefore its resistance in the divider) can be changed by varying the CR2 cathode potential. The voltage across C13 is normally equal to the CR2-R23 potential, and, when signals are being received, C13 attempts to follow this voltage. Thus, when a channel pulse from the V4 plate arrives at the CR2 cathode, the CR2-R23 voltage decreases. C13 discharges to this voltage through R23 and the conducting resistance of CR2. One time constant of the C13 discharge period is approximately 0.1683 μs. Therefore, the leading edge of the pulse at the V5 grid is a sharp negative input.

At the end of the negative channel pulse, C13 must charge to its quiescent voltage through R24. One time constant of this circuit is approximately 5.940 μs.

When the leading edge of the marker pulse group is applied to the CR2 cathode, the same action occurs. However, the 0.8-μs spacing between marker pulses is not sufficient to allow C13 to recharge to its quiescent voltage. Thus, the marker (or sync) pulse is identified at the V5 grid as a wider pulse, having its width slightly greater than that of the total marker pulse.

Sync separator V5, separates the signal pulses from the rest of the video frame. V5 is normally conducting as a result of a grid connection through R24 to the B supply. Each of

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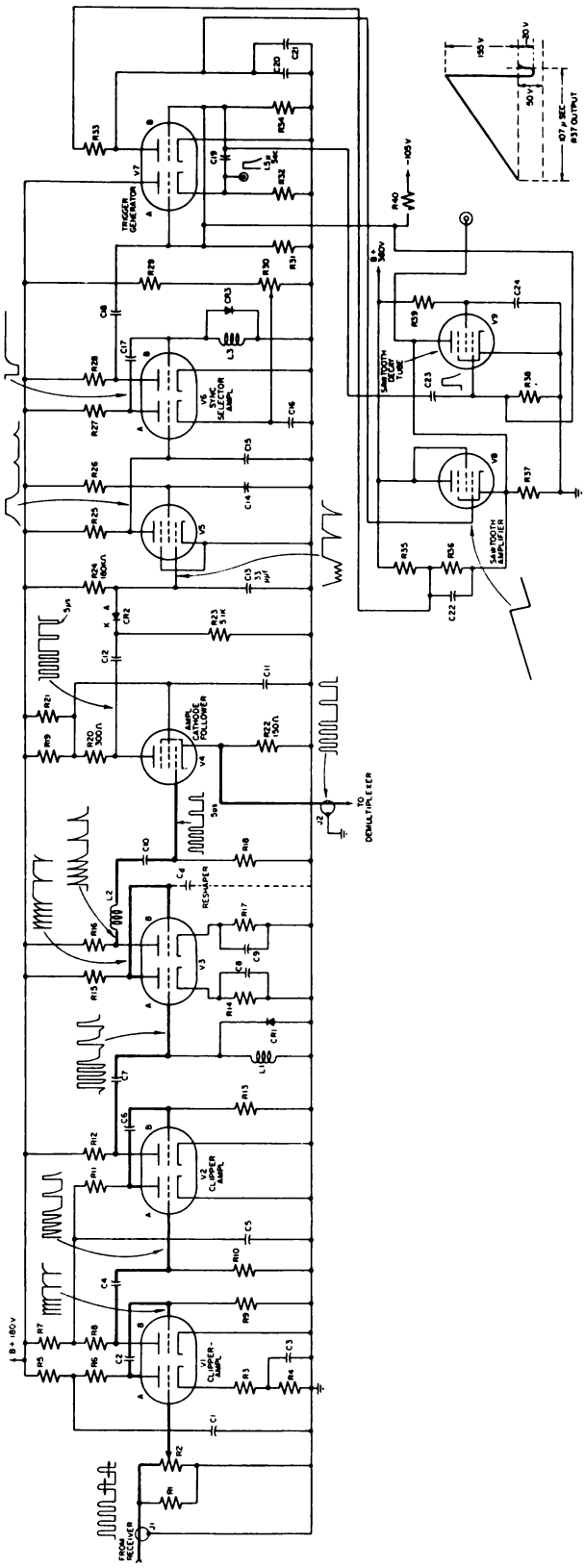

Figure 9-12. —Demultiplex waveform generator.

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the negative input pulses are sufficient in amplitude to drive V5 beyond cutoff. The plate voltage rise charges C15 through R25. One time constant of this circuit (RC) equals about 5.280 μs. Therefore, the amplitude of the channel pulses (duration approximately 0.5 μs) developed by C15 is small, relative to the amplitude of the wider marker pulse during their respective integrating periods.

Sync selector amplifier V6 is normally held beyond cutoff bias by a positive voltage at its cathode obtained from a voltage divider comprising R29 and R30. The bias may be varied by adjusting the arm of R30. With the proper bias setting, only the integrated marker pulse (described above), which appears at the grid, should be of sufficient amplitude to trigger V6A into conduction. The resulting plate voltage drop produces a negative-going pulse, which is shaped by L3 to approximately 1.5 μs. Periodic CR3 conduction damps positive alternations developed by L3. The negative pulse developed by V6A represents the trigger pulse, which will be used (as described later) to initiate the generation of the demultiplex sawtooth waveform.

The positive sawtooth trigger pulse output of V6B is coupled by C18 to the trigger generator, V7A, control grid. The A section of V7 functions as a biased cathode follower, which is triggered by the V6B input to produce a positive sawtooth trigger output. (The sawtooth designation refers to the waveshape of the final output of the demultiplex waveform generator and not to the waveform of the trigger itself.)

When the B supply voltage is initially applied C20 and C21 in the V7B plate circuit begin charging through R33 toward the B supply value. The positive potential developed by these capacitors is applied to the control grid of the sawtooth amplifier, V8. The positive-going grid potential causes increased current through R37. The increasing voltage developed by R37 is applied through C22 (as a transient voltage) to the R33-R35 junction, effectively increasing the potential to which C20 and C21 must charge. The increased charge of these capacitors causes a more positive voltage at the V8 grid, and the R37 voltage rises to still a higher value. Thus, a constantly rising potential is applied through C22 to the R33-R35 junction to cause the sawtooth boot-strap rise in the charge of C20 and C21.

It should be noted that the trailing edge of the sawtooth trigger pulse developed by V7A occurs at the same time as the trailing edge of the sawtooth trigger pulse in the multiplex waveform generator. This pulse is applied along dual paths by C19 to the V7B control grid, and by C23 to the V9 control grid. Both of these tubes received a cutoff bias potential from a -105-v source (not shown). V7B conducts to discharge C20 and C21, while V9 conducts to short R37, causing the sawtooth decay.

Since the trailing edge of the multiplex and demultiplex sawtooth trigger pulses occur at the same time, the sawtooth linear rise of the two systems will also occur simultaneously, and the synchronism between the two systems is assured.

Channel Demodulator

The sawtooth input from the demultiplex waveform generator (fig. 9-12) is applied through R50 (fig. 9-13) to the grid of the demodulator channel selector, V10. This circuit functions in the same manner as described for the channel selector in the modulator. Bias is obtained from the arm of R42, which forms a portion of a voltage divider comprising R41 through R48, and connected from the B supply to ground. Only the circuit for the channel 6 demodulator is shown in the figure. All channel demodulators are identical in circuitry. They differ, however, in the bias level applied at their respective cathodes, and likewise in the time the circuit demodulating process is initiated.

When the sawtooth input rises sufficiently in amplitude to overcome the V10 bias, the plate voltage drop of V10 is differentiated across the long time-constant circuit comprising C26 and R53 to produce a wide negative pulse at the V11 grid. The trigger stage, V11, normally conducts since it is cathode biased. Upon arrival of the negative input at the V11 grid, the V11 plate voltage rises to charge C28 through R56 and C30.

V12A is normally held cut off by a fixed bias voltage obtained from the -17-v supply. When the positive input appears at the grid, V12A conducts. A negative trigger pulse is applied through C31 to the T1 primary.

The positive secondary potential of T1 charges C32 through R58 to cause a positive input to be applied to the V12B grid. V12B conducts a heavier current, which causes a drop in its plate voltage. The drop appears as a more negative signal in the T1 primary, which is coupled to the secondary of T1 as a more positive potential at the V12B grid. The grid conducts a

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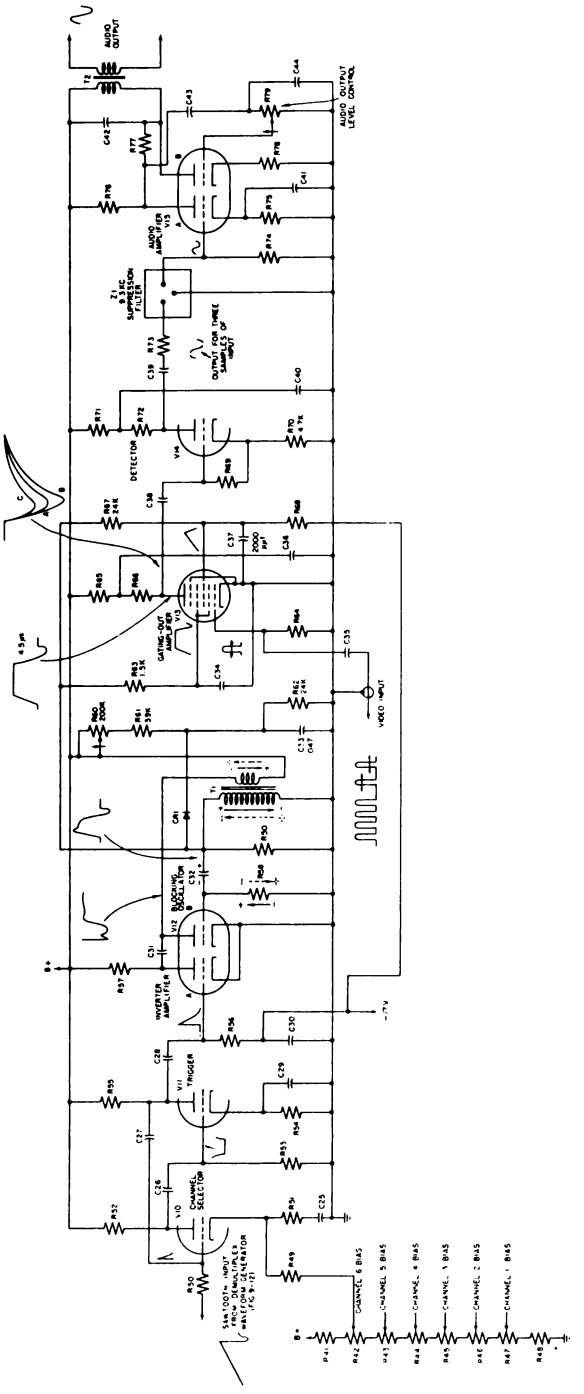

Figure 9-13. -Channel demodulator, schematic diagram.

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current to charge C32 negative toward the grid. The regenerative action is cumulative until V12B conducts saturation current.

At saturation, no further drop in the V12B plate occurs, and the positive voltage at the T1 secondary decreases to zero. C32 now discharges through R58, applying a high negative potential to the V12B grid to cut off this section of the tube. The plate voltage immediately rises to the B supply value, which is aided momentarily by the collapsing field of the T1 primary. This action induces a small negative voltage at the T1 secondary, which aids in the V12B cutoff, and is responsible for the negative overshoot of the output pulse.

CR1 normally does not conduct since its cathode receives a positive bias potential from the R61-R62 junction. R62 (paralleled by C33), R61, and R60 form a voltage divider from B plus to ground. If the arm of R60 is turned clockwise (as indicated by the direction of the arrow) a greater portion of the total resistance is in the circuit, and the drop from R62 to ground is smaller. This smaller positive potential is applied to the CR1 cathode, and the diode will conduct with less positive voltage at its anode. As the R60 arm is turned counterclockwise (opposite to the direction of the arrow) more of the total resistance is removed from the divider, and the B supply voltage is divided primarily between R61 and R62. The potential across R62 is increased, and the CR1 cathode is more positive.

The amplitude of the blocking oscillator (V12) output pulse is approximately 40 v. The pulse width is adjusted to 4.5 μs. The width of the pulse varies somewhat sinusoidally as the voltage rises due to the oscillatory action of the T1 secondary. If the R61-R62 junction potential is adjusted to, say 35 v by R60, CR1 will conduct through the low-charging reactance of C33 when the anode potential exceeds 35 v. The resulting output pulse width of the blocking oscillator, which is coupled through R63 and R67 to the respective V13 grids, is thereby limited to the width of the pulse (along its rise) at the time of the CR1 conduction. CR1 shorts R59 by way of C33. If R60 is adjusted clockwise, the CR1 cathode potential becomes smaller, and limiting will occur earlier. This results in a wider output pulse to V13. Conversely, if R60 is adjusted counterclockwise, the R61-R62 junction potential increases and CR1 conducts later along the rise time of the blocking oscillator output. The result in this case is a narrower pulse at the output.

Since the sawtooth waveform of the multiplex and demultiplex systems are synchronized, the occurrence in time of the trigger to the channel 6 demodulator is coincident with the time of arrival of the channel 6 modulated pulse in the video frame. The total video frame is applied at the V13 control grid.

V13 is a coincidence stage, which requires the simultaneous application of four potentials before conduction will occur. The output of the blocking oscillator coupled through the relatively short time-constant circuit comprising R63 and C34 is not altered in shape, and serves as the screen potential for V13. The same blocking oscillator input is coupled through a long time-constant circuit comprising R67 and C37. The charge of C37 is relatively slow in this circuit, resulting in a sawtooth (integrated) waveform at the injector grid. The channel pulse occurs at the control grid as discussed earlier. The B supply voltage is constantly applied at the plate. A small negative (bias) voltage is applied at the injector grid, which is overcome by the rising sawtooth voltage to allow the tube to conduct if the other potentials are applied.

If the channel pulse is unmodulated, it will be positioned (in time) toward the center of the 4.5-μs gate at the screen grid, and at about the midamplitude point along the rise of the sawtooth waveform. The tube conduction will result in a drop as shown at "a" (fig. 9-13) for the plate waveform. If the channel pulse is modulated and at the right of center on the sawtooth rise, the tube conducts heavier, and the plate potential is lower as represented at "b". Conversely, if the modulated pulse appears to the left of center along the sawtooth amplitude rise, V13 conducts a smaller current, and the plate voltage decrease is correspondingly less as shown at "c." Thus, the position of the pulse in the video frame determines the amplitude of the output pulse from V13.

It should be understood that only one pulse will appear at the V13 plate for each video frame (fig. 9-9). Since the frame recurrent frequency is 9.3 kc, the amplitude of the V13 output varies at this rate (a to b to c if the channel is modulated).

The negative output of V13 is coupled by C38 to the control grid of the detector, V14. The stage is operated at zero bias, and the cathode is held above ground by the amount of the voltage developed across R70. That portion of the V13

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output appearing across R69 is amplified in the plate circuit and contains both the modulation component and the 9.3 kc-sampling frequency component of the multiplex signal. The leading edge of the V13 output develops operating bias across R69, which moves the operations of V14 to a lower point on the plate current-grid voltage curve. In this manner, only the range in which the amplitude modulation exists in the V13 output is detected.

The V14 output is coupled by C39 and R73 to a 9.3-kc suppression filter, Z1. The filter allows easy passage of the lower audio frequencies ranging from approximately 300 to 3,000 cps and blocks the passage of the 9.3-kc sampling frequency.

The Z1 output is amplified by V15A and developed across C43 and R79. The setting of the arm of R79 determines the amplitude of the audio input to the V15B grid. This input is amplified and applied across T2 as the audio output.

TIME-DIVISION MULTIPLEX USING A DELAY LINE

In any time-division multiplex system, the development of an accurate method of energizing the various channels at a given time is of utmost importance. The system must employ some type of highly stabilized master oscillator. In the system just discussed, a triggered boot-strap sawtooth oscillator was used in both the multiplex and demultiplex equipments. A crystal-controlled oscillator would serve equally as well.

A block diagram of a time-division multiplex system using a crystal-controlled master oscillator and delay line is shown in figure 9-14. The oscillator generates an 8-kc sine wave, which is acted upon (clipped) by the pulse generator

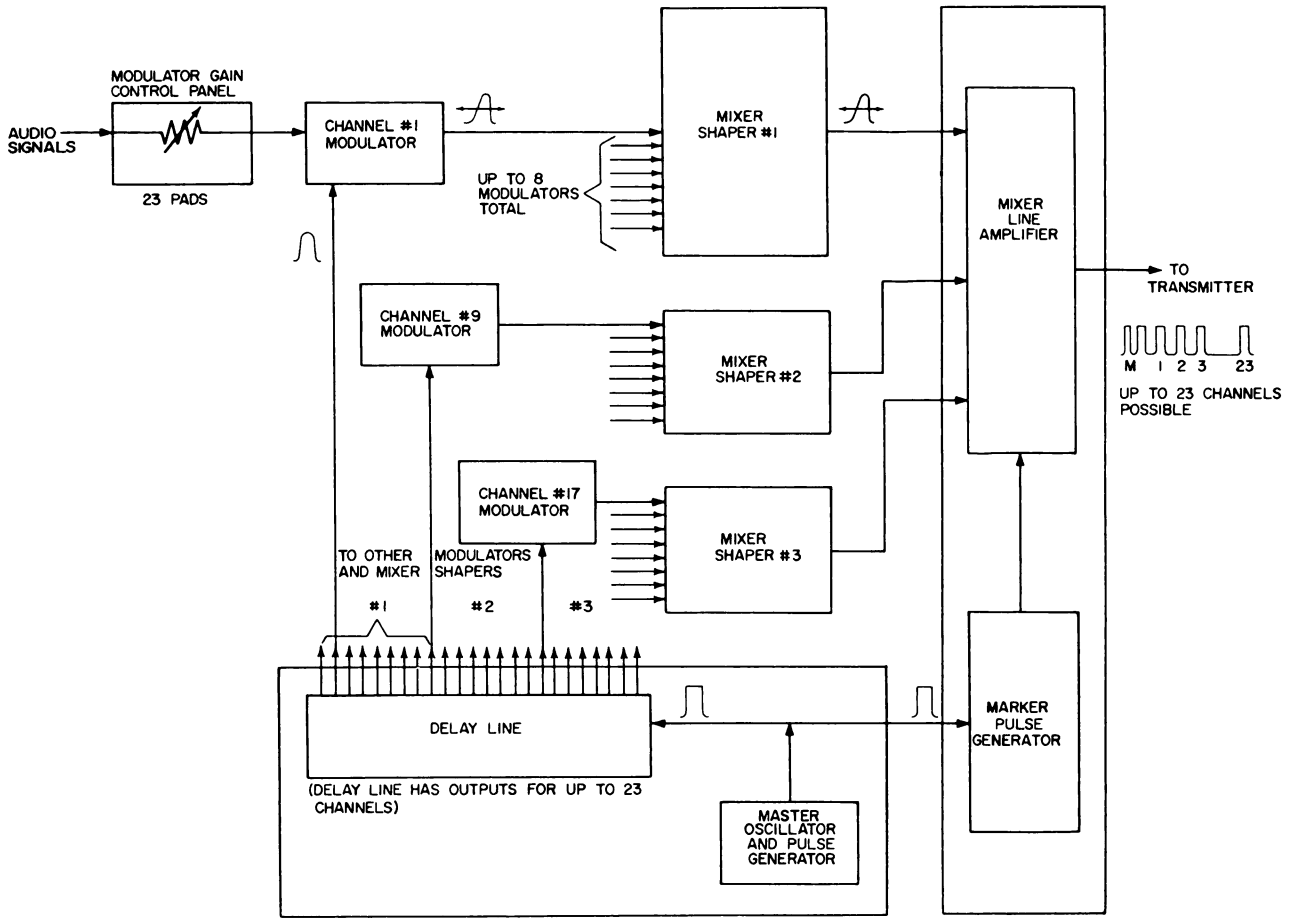

Figure 9-14. —Block diagram of multiplex modulation circuits.

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circuit to produce the electrical equivalent of a square wave. The oscillator frequency (8 kc) is the frame or base frequency of the multiplex system. This output is coupled along dual paths to a marker pulse generator, and to the delay line.

The artificial delay line comprises 24 sections of pi-connected inductors and capacitors (not shown), each of which introduces a 5.2- μs delay to the passage of the pulse along the line. A connection is made to a modulator channel from each section of the delay line. Thus, 23 channel modulators can be activated at time intervals of 5.2 μs by the trigger pulses from the line. The presence of the controlling pulse enables the modulator to generate one channel pulse per frame (8,000 per second). The audio waveform applied to the modulator varies the position of the pulse in a manner similar to that discussed earlier. During the 24th part of the frame interval, a double-pulsed marker is produced for synchronizing the multiplex system.

To minimize crosstalk, the modulators are arranged in three groups of eight. The output of each group of eight modulators is combined in the mixer shaper, which shapes each modulator output into a narrow pulse. The output of each of the mixer-shaper stages is applied to a single mixer line amplifier, which combines all the channel pulses and marker pulses into a single video frame. The mixer line amplifier output is applied to the r-f transmitter circuits.

The pulse train input to the demodulator (fig. 9-15) comprises the marker, and up to 23-position (time) modulated channel pulses. The signals are shaped in a clipper-shaper stage and applied to the marker-separator.

The marker-separator circuit extracts (detects) the double-pulsed marker and converts it to a suitable square waveform, which is fed to a 24-section delay line. This delay line is identical to that employed in the multiplex circuits.

The line produces the controlling pulses (properly spaced in time sequence) for activating the individual channel demodulators. The first output pulse is delayed 5.2 μs from the input pulse. Each of the remaining pulses is delayed successively 5.2 μs from the preceeding pulse.

The first 23 pulses from the delay line form the control or deblocking pulses for the 23 demodulators. Each demodulator employs a coincidence stage (not shown), which requires the simultaneous application of the channel pulse and the delay line deblocking pulse to cause conduction. The entire video frame is applied to each demodulator. Since the deblocking pulses from the delay line are applied at specific intervals, only one demodulator will be activated for a single interval. The demodulated signals are filtered and amplified and reproduced at the output.

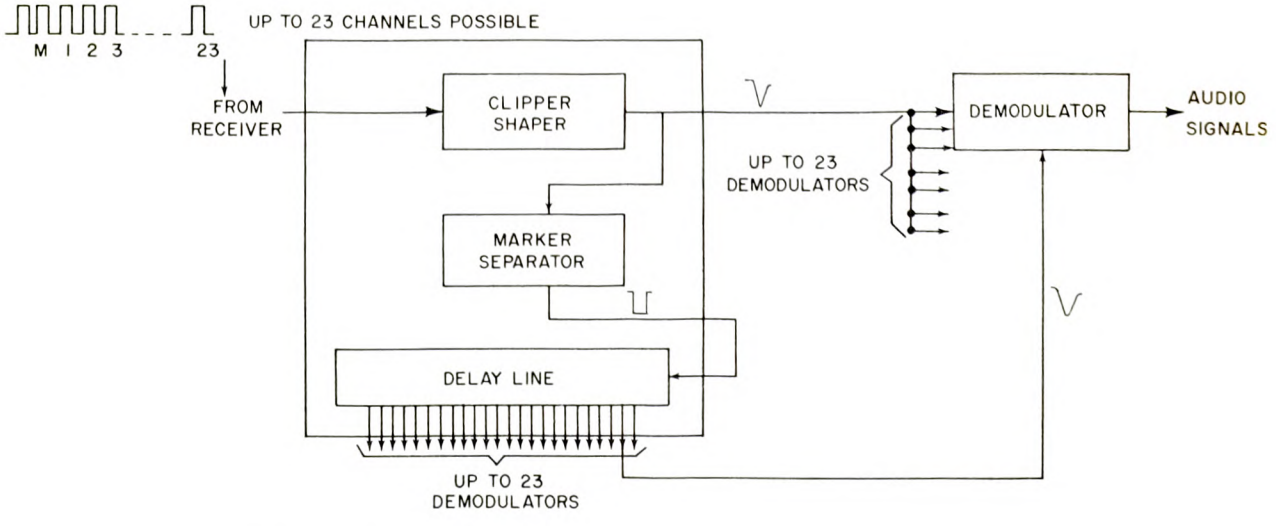

Figure 9-15. –Block diagram of multiplex demodulation circuits.

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QUIZ

1. The transmission of several intelligible signals during the same period of time normally required for the transmission of a single signal is called what type of transmission?
2. To use time in the most efficient manner, in time-division multiplexing, what would be the bandwidth necessary to transmit n channels?
3. What is the total bandwidth required for a frequency-division multiplexing system?
4. What are the three practical methods of pulse modulation?
5. If the rate of sampling throughout the multiplex system is adjusted to 9.3 kc, what will be the maximum allowable audio frequency that may be used in the system?
6. The duration of the sawtooth waveform (fig. 9-6) is adjusted to exactly what value?
7. The sawtooth waveform (fig. 9-6) is fed along paralleled paths to what circuits?
8. The multivibrator output trigger (fig. 9-6), which is fed to the sawtooth oscillator and discharge tubes, is of what duration?
9. What is the primary purpose of the marker pulses (fig. 9-6)?
10. When the V6B plate voltage rises (fig. 9-6), a positive 4.75-μs pulse is developed at the grid of what stages?
11. The negative portion of V6 output (fig. 9-6) is applied through C17 to the control grid of what stage?
12. The R76 arm (fig. 9-7) should be adjusted so that the charge time of C30 necessary to overcome the V17A bias is of what duration?
13. The amplitude of the audio signal at the time of the channel pulse determines the magnitude of the V17A bias voltage (fig. 9-7) which must be overcome by the charge of what capacitor?
14. If the signal input to the delay line (fig. 9-8) is unmodulated the position of the leading edge of the negative (reflected) pulse will always be at what position in time from the leading edge of the input pulse?
15. The successful reproduction of the signals in the demultiplexer depends primarily upon what feature of the system?
16. The trailing edge of the sawtooth trigger pulse developed by V7A (fig. 9-12) occurs at the same time as the trailing edge of what other pulse in the system?
17. The amplitude of the blocking oscillator (V12) output (fig. 9-13) is approximately what value?
18. Since the sawtooth waveform of the multiplex and demultiplex systems are synchronized, the occurrence in time of the trigger to the channel 6 demodulator is coincident with the time arrival of what pulse in the video frame?
19. What feature of the channel pulse in the video frame (fig. 9-13) determines the amplitude of the output pulse from V13?
20. The artificial delay line (fig. 9-14) comprises 24 pi-connected inductors and capacitors (not shown) each of which introduces what amount of delay to the passage of the pulse along the line?

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CHAPTER 10

RADIO TELETYPE SYSTEM, PART I

INTRODUCTION

A radio teletype system (fig. 10-1) is a series (loop) connected telegraphic communications network operating between two or more points, and involving the transmission and reception of coded signals. A brief discussion of the basic principles of teletype operation is given in this chapter. It is assumed that you are already familiar with the information on teletype systems contained in *Electronic Technician 3*, NavPers 10188.

BLOCK DIAGRAM OF A REPRESENTATIVE RADIO TELETYPE SYSTEM

A block diagram of a representative radio teletype system is shown in figure 10-1, A. The specific equipments in the diagram are discussed in this chapter.

Before discussing the purpose of the individual units, note that all major components in the transmitting and receiving teletype terminal equipments are series connected, depending on the setting of the send-receiver switch in the control unit.

The receiving series loop includes the page printer, power supply, teletype panel, converter-comparator, and the control unit. In this case, the send-receive switch in the control unit is set in the receive position. During frequency-shift keying the transmitting, or send loop, includes the page printer, power supply, teletype panel, frequency-shift keyer, and control unit with the switch in the position shown. When the

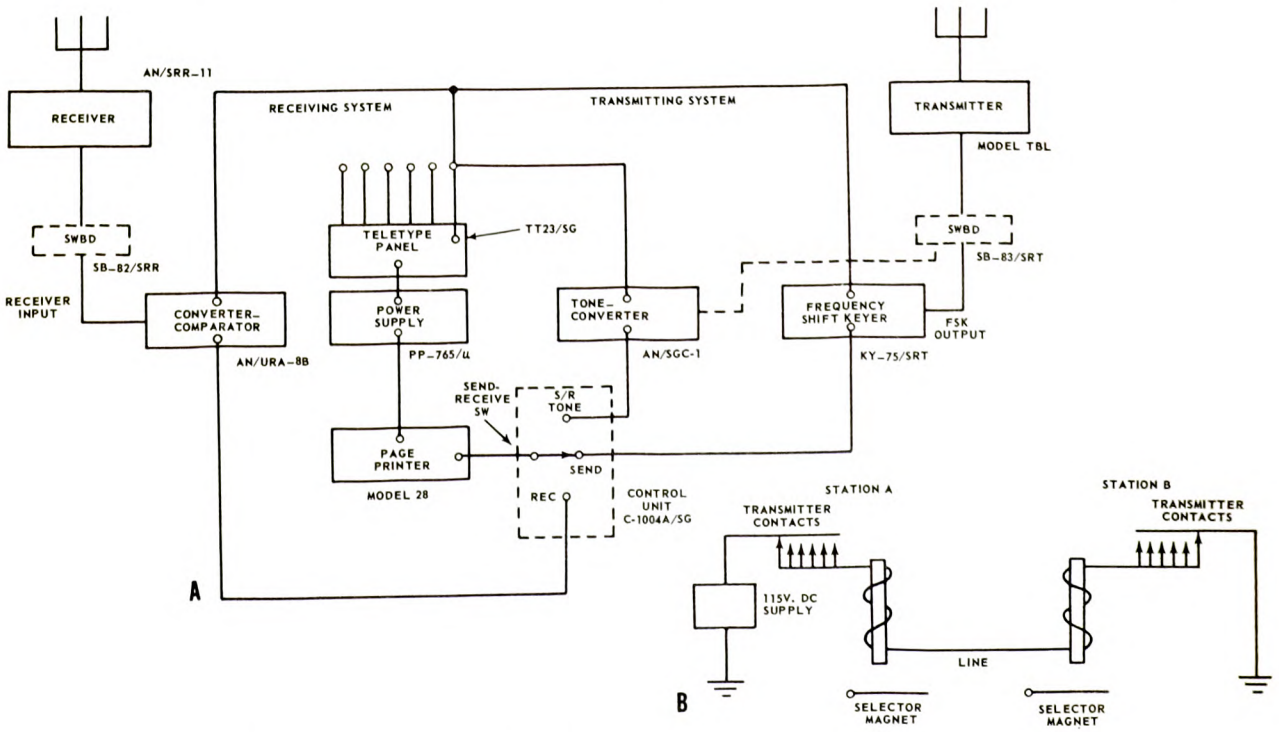

Figure 10-1. —Block diagram of a representative radio teletype system.

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switch in the control unit is set to the S/R TONE position, the transmitting loop is completed through the tone converter.

A single power supply, PP-765/U (picture not shown) provides a current to the chosen series loop, which operates the loop components. The power output of the source is rated at 120 volts d-c (±1.5 volts) from 0 to 1.0 ampere. Six teletypewriter systems may be operated from the power source through proper connection to the teletype panel. During transmission the page printer (not shown) opens and closes the teletype loop, thereby modulating the line current with a different set of a five-unit code for each letter, figure, or symbol to be transmitted. During reception, the page printer reproduces or prints the message sent from the remote radio teletype system.

The five-unit code is composed of five selecting intervals, which may be either marking impulses, causing current in the line or spacing impulses in which no line current flows. Each group of five selecting intervals is preceded by a start interval, or space (no current). A stop (or mark current) impulse follows each five-unit code. Both the start and stop impulses are used to maintain synchronism between the transmitting and receiving systems.

The teletype panel, TT-23/SG (fig. 10-4) contains six jacks, which make possible the interconnection and transfer of the teletype equipment. Each system to be operated is patched through the teletype panel. A selector switch, rheostat, and meter on the front panel are provided for adjusting the loop current in any channel.

The frequency-shift converter, KY-75/SRT (fig. 10-6) may be used with many different types of Navy transmitters, including such types as the TBL and TBK. When the page printer keyboard produces a marking signal, the output from the KY-75/SRT causes the transmitter to emit a frequency above its assigned or rest frequency. Conversely, when a space signal is produced at the page printer keyboard, the transmitter emits a frequency below its normally assigned frequency.

The frequency-shift keyer discussed here transmits a type F1 frequency-shift signal in the frequency range from 1 to 6.7 megacycles. The output of the keyer can be applied to any c-w transmitter capable of operating from a 2- to 20-volt excitation source. The signal is passed through the class "C" amplifiers (or multipliers) of the transmitter to the antenna.

The control unit, C-1004A/SG (fig. 10-11) is provided to switch the teletypewriter to any one of three d-c loops. The three channel loops include (1) a tone send-receive terminal, wherein the tone converter is connected in the loop, (2) a frequency-shift terminal, which connects the frequency-shift keyer in the loop, or (3) a receiver circuit, which incorporates either a frequency-shift converter or converter-comparator, both of which perform primarily the same function.

The tone converter, AN/SGC-1 (figure 6 of chapter 11) provides another method of transmitting the teletype information. This method involves the transmission of two audio tones, one of which corresponds to a mark while the other produces a space.

A single position of the switch on the control unit (S/R TONE) serves for both transmission and reception of the intelligence when the tone converter is in use. The circuits of the tone converter are such that they close the line to the transmitter (which places the unit on the air) at the instant the operator begins typing the message. The line remains closed until after the message has been transmitted.

During reception of either frequency-shift or tone signals, the incoming mark and space frequencies are acted upon by the receiver to produce audio tones. The tones are then converted to ON and OFF d-c pulses in the converter-comparator (in the case of frequency shift signals) or in the tone converter (when the signals are being received). The output pulses of the converter-comparator make and break the teletypewriter direct current loop, which causes the page printer to reproduce the message.

The converter-comparator, AN/URA-8B, (figure 1 of chapter 11) is designed primarily for use during diversity reception. This unit contains a comparator (used during diversity reception) and two frequency-shift converters, which convert the pulses from the receiver as discussed. The frequency-shift converters may be used separately for single-receiver reception of the frequency-shift signals. Each converter has its own output circuits for keying the teletype d-c loop and for providing a keyed audio tone. When the converters are used in single-receiver reception, the comparator may be associated with one receiver to provide an additional output circuit.

A simplified schematic of a basic teletypewriter land-line network is shown in figure 10-1,B. If station A is transmitting, the five

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transmitter contacts are positioned in accordance with the character to be transmitted. A marking impulse causes current to flow through both the station A and station B selector magnets. The selector magnets, in turn, cause the page printer to reproduce the character both locally and at the distant station B.

The operation of station B (in transmitting) is identical to that of station A. The message may be transmitted over either line or radio facilities.

PAGE PRINTER (MODEL 28)

MOTOR

A single motor is provided to drive the mechanism of the page printer. The motor used with the Model 28 page printer (fig. 10-2) is designed to operate from a single-phase, 115-volt, 60-cycle, a-c supply. It is a .05-horsepower, 3600-rpm, two-pole, wound-stator, ball-bearing motor with a squirrel cage rotor. The stator windings include a main operating winding and a starting winding.

The starting winding is connected in series with a 43- to 48-μf electrolytic capacitor, which produces split-phase currents as required for starting the motor. A centrifugal switch, S2, is connected in series with the capacitor. S2 is held closed by a spring (not shown) when the motor is stopped, and remains closed from the

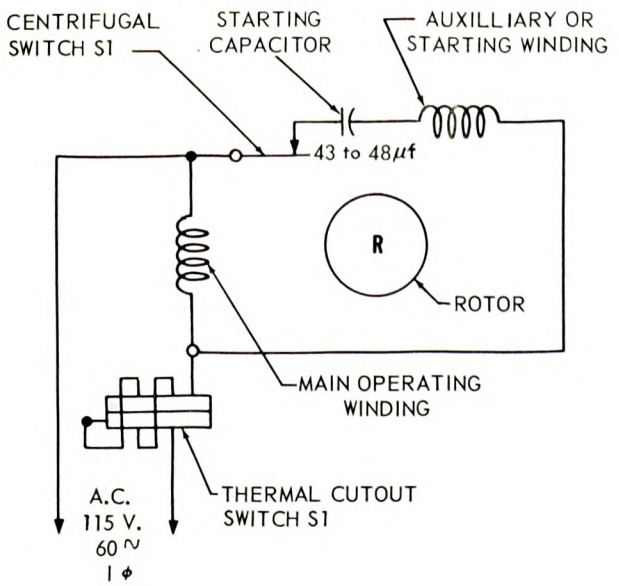

Figure 10-2. —Model 28 page printer motor, schematic diagram.

time the motor is started until the rotor acquires a speed of approximately 2700 rpm. At this speed, the centrifugal force developed by the motor is sufficient to overcome the tension of the spring so that S2 is opened. The opening of S2 disconnects the auxiliary, or starting winding, and its series capacitor from the line voltage. The rotor continues to accelerate until it reaches its synchronous speed of 3600 rpm.

To prevent overheating and possible damage to the motor if it is stalled, a thermal cutout switch, S1, is installed in series with both the main and auxiliary motor windings. If a current greater than 11 amperes is drawn by the motor, S1 will open the line voltage circuit. A manual reset button (not shown) may be used to close S1 when circuit and load conditions permit.

KEYBOARD

The motor is geared to the main shaft of the typing unit, which, in turn, extends motive force to the keyboard mechanism. Code signals are applied to a two-coil magnet (not shown), which is associated with a selecting mechanism. The mechanism interprets the signals and controls the mechanical motions involved in typing a character.

A detailed discussion of the keyboard is not included in this chapter since it is largely mechanical. Its maintenance and repair are not the responsibility of the Electronics Technician. The Model 28 keyboard mechanism operation can be found in *Teletypewriters Instruction Book*, NavShips 91393.

POWER SUPPLY PP-765/U

Input power to the power supply from a 115- or 230-volt, 50-60-cycle, single-phase, a-c source is applied to terminals A and C of J1 (fig. 10-3). ON-OFF switch S1 and fuses F1 and F2 complete the circuit through the connections at the terminal board, TB1, to the primary of the power transformer, T1.

When a 115-volt input is used, the two primary windings are connected in parallel. When a 230-volt input is used, the windings are connected in series.

Rectification of the output voltage is accomplished by two selenium rectifiers, CR1 and CR2. The rectifiers are connected in a single-phase, full-wave circuit. The a-c input voltage is applied to the rectifiers through

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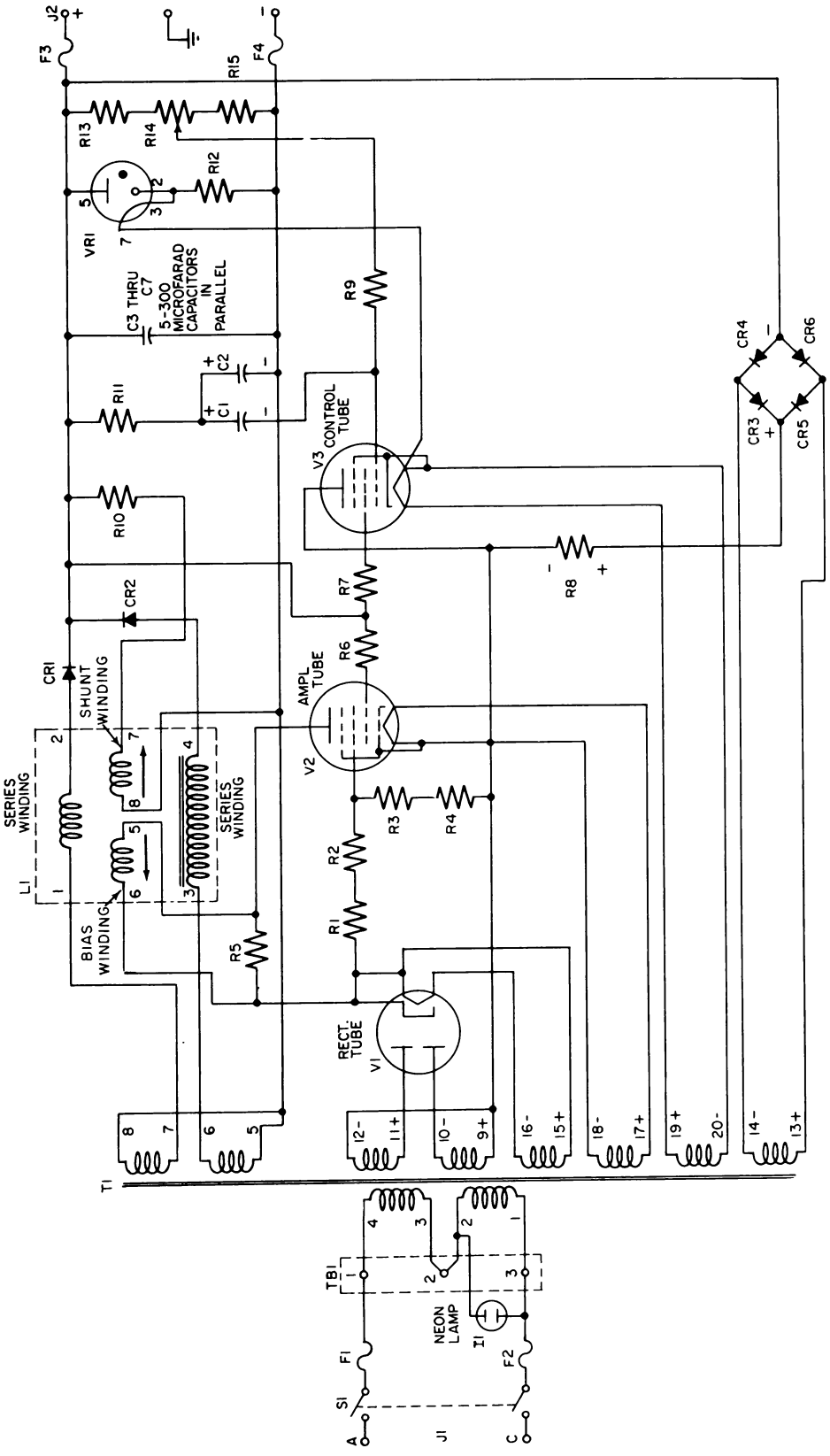

Figure 10-3. —Power supply PP-765/U.

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terminals 8-7 and 6-5 of T1. C3 through C7, in conjunction with reactor L1, (discussed later) filter the output voltage to reduce the ripple content to less than 1 volt rms, or .83% of the output voltage.

The action of the power supply is as follows: Assume that there is a positive potential at terminal 7 of the T1 secondary. Electron flow is then from terminal 8 through the filter capacitors (and all of the load components) through CR1 and L1 (terminals 2 and 1) and back to the positive terminal. The conduction path is the same during the next half cycle of the a-c input, except that CR2 conducts through L1 (terminals 4 and 3).

L1 is a saturable swinging choke, which comprises four sections. In conjunction with the other filter components, it regulates the output voltage of the supply.

Coils 1-2 and 3-4 are called series windings since they are connected in series with the power supply circuit. Coil 7-8 is connected in series with R10, and the combination of the two components is connected in shunt with the output. This winding is called the shunt winding. Coil 5-6 controls the magnitude of the correction permitted in the output voltage variations (as discussed in the following paragraphs) and is called the control, or bias, winding.

A certain value of the output voltage developed across the voltage divider comprising R13, R14, and R15 is taken from the arm of R14 and applied through limiting resistor R9 to the control grid of V3. If the load current is increased, the voltage across the divider tends to decrease. Conversely, if the load decreases, the voltage output tends to increase. The potential at the arm of R14 varies in accordance with the load on the supply.

Assume that there is a load decrease with its corresponding output voltage increase. The voltage at the arm of R14 and, likewise, the control grid potential of V3 increases. However, because R13, R14, and R15 form a voltage divider across the output, only a portion of the full voltage change will appear at the arm of R14.

The voltage across VR1 will remain constant since VR1 is a voltage regulator tube, and the entire change of voltage will appear across R12. Thus, while the potential at the control grid of V3 has increased, the potential at the cathode of V3 (connected to the cathode of VR1) has increased even more. This causes the conduction of V3 and the voltage drop across R8 to decrease, and the plate potential of V3 to increase by an amount even greater than the change developed across R12.

The positive going potential at the plate of V3 is applied to the cathode of V2. This positive change in potential at the V2 cathode is greater than the positive change that takes place at the control grid of V2 through R6. Thus, the bias on V2 is increased, causing V2 to conduct less to decrease the conduction through the 5-6 bias coil.

When the load decreases, the 7-8 shunt coil current and, likewise, the 7-8 winding magnetomotive force increase slightly. The shunt coil is wound so that the 7-8 winding magnetomotive force is in opposition to the magnetomotive forces of the 1-2 and 3-4 series winding and of the bias 5-6 winding.

Because the 5-6 winding current has decreased as a result of a smaller conduction of V2, the 5-6 winding bias is also decreased, permitting the 7-8 shunt winding magnetomotive force to further increase. This action causes decreased saturation in the 1-2 and 3-4 windings of L1, which is accompanied by an increase in the impedance of these windings. Since this impedance appears in series with the load connected between J2 and ground, the tendency of an output voltage rise across the load is counteracted.

The plate potential of V3 is approximately 87 volts. This value of voltage is derived through the action of the voltage regulator tube, VR1, and the bridge circuit comprising CR3, CR4, CR5, and CR6. The VR tube supplies a relatively fixed voltage between the cathode and plate of V3 of approximately 75 volts. An a-c input derived from terminals 13 and 14 of the power transformer is rectified by the bridge circuit.

Assume that there is an input voltage at the T1 secondary such that terminal 13 is positive and terminal 14 negative. Electron flow is then through CR4, through one of the rectifiers (CR1) of the power supply, through L1 (terminals 1-2) and the conducting half of the T1 secondary (terminals 8-7), through the center tap, through R12 and pins 3 and 7 of the VR tube, through the cathode-to-plate resistance of V3, through R8, and through CR5 to terminal 13 of the transformer. The next half cycle electron flow is through CR6, CR2, coil 4-3 of L1, coil 6-5 of T1, R12, terminals 3 and 7 of VR1, the cathode-to-plate resistance of V3, R8, CR3, and terminal 14 of T1. The output voltage of the bridge rectifier aids the voltage supplied by the VR tube,

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thus boosting the potential at the V3 plate to approximately 87 volts.

R11 (fig. 10-3) is a feedback resistor, which functions in conjunction with C1 to supply a small amount of the output voltage change to the V3 control grid. C2, in conjunction with R11, feeds back sudden changes in the output voltage that bypass the VR tube. Both actions tend to prevent an oscillating or hunting effect by the power supply.

Operating voltage for V2 is obtained from a separate power supply, which employs V1 as the rectifier tube and terminals 9 through 12 of T1 as the power transformer. The circuit is connected as a conventional full-wave rectifier.

TELETYPE PANEL TT-23/SG

The teletype panel TT-23/SG (fig. 10-4) is intended for general shipboard use to provide the interconnection and transfer of teletype equipment with various radio adapters. The term, radio adapters, includes such units as frequency-shift keyers and converters and tone modulation terminal equipments.

The panel contains six "looping" circuits. The word, looping, is intended to indicate the manner in which the jacks are connected in series with the teletype equipment and radio adapters, thus forming a complete loop.

Each looping circuit of the panel comprises two looping jacks (fig. 10-5, B) one set jack, a rheostat for adjusting line current, and a toggle switch, S1, for selecting either a local or external source of line current.

The meter, M, and the meter selector switch, S2, are provided for measuring the line current in either of the six channel loops. The meter shunt for each circuit has the same resistance value. Thus the meter may be switched from one channel to another without interrupting the teletype signal.

With the jumper connection of the looping jacks as shown, the tip and sleeve springs are shorted through the spring contacts, thus providing circuit continuity. When a patchcord plug (from the page printer or other equipment) is inserted in either of the looping jacks, the spring contacts open, allowing the equipment to be connected in the series circuit.

All teletype equipment desired to be connected in the looping circuit should be connected to terminals 3 and 4. Terminals 3 and 4 in figure 10-5, A, correspond to the right digit of the two-digit number. The first digit represents the channel number. All radio adapters, such as

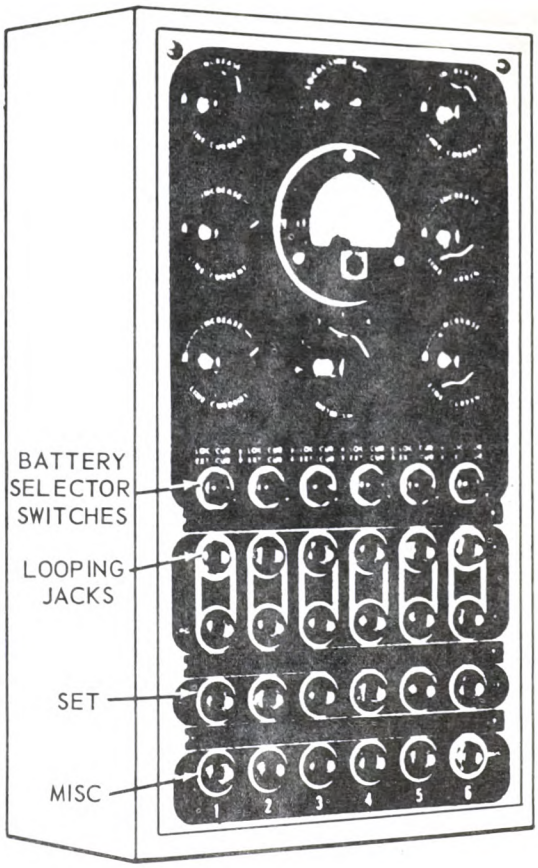

Figure 10-4. —Teletype panel TT-23/SG.

frequency-shift keyers, converters, and tone modulated terminal equipment, should be connected to the 1 and 2 terminals. Other miscellaneous teletype equipment, such as reperforators and transmitter distributors, should be connected to terminals 5 and 6.

The "set" jack incorporates a double-pole, double-throw switch. When no plug is inserted in the jack, it functions in the same manner as the looping jacks, and the teletype equipment connected at terminals 3 and 4 is placed in the loop.

The insertion of a plug in the "set" jack (fig. 10-5, C) opens the circuit from the teletype equipment to the looping jacks. In addition, the looping circuit of the channel containing the "set" jack is closed, retaining continuity through the looping jacks and the radio adapters of that channel.

If it is desired to transfer a teletypewriter from one channel to any other channel, patch the teletype equipment from the corresponding "set" jack to one of the looping jacks in the

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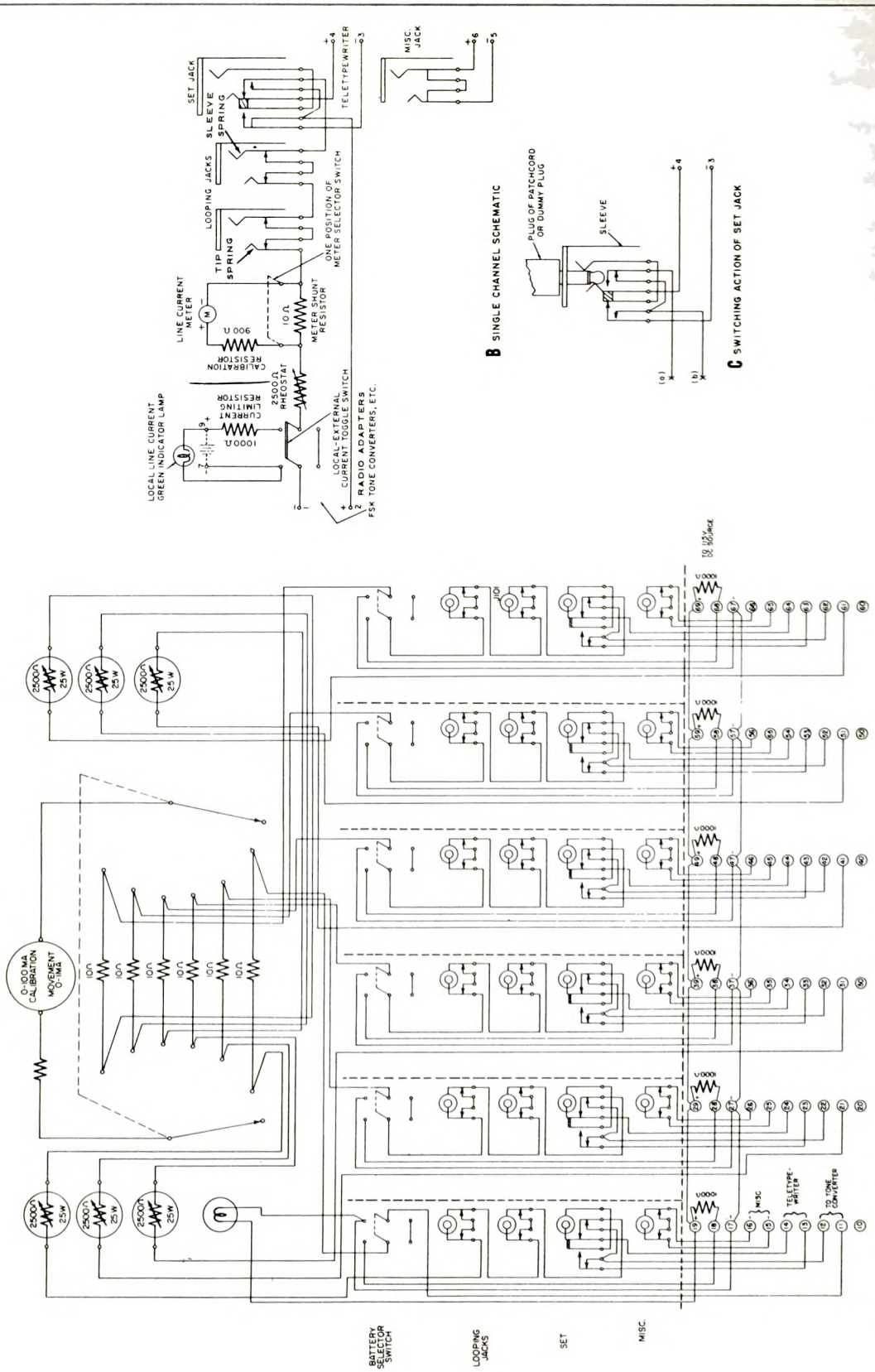

Figure 10-5. —Teletype panel, schematic diagram.

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channel to be used. The loop is then completed through the radio adapters of the new channel, receiving the operating power from that channel. If another teletypewriter is patched from a second "set" jack to the unused looping jack of the same channel, two teletypewriters may be used (alternately) to operate the same radio adapters.

If it is preferred that the teletypewriter wired in any particular channel should not operate, insert a dummy plug in the "set" jack of that channel. The six miscellaneous jacks are provided as a connection for any teletype equipment not regularly assigned to a particular channel.

Provisions are made to connect a local source of 115 volts d-c across terminals 7 and 9. The connections are made such that they parallel the corresponding terminals of each channel. The local current can be switched in and out of each channel by means of the selector toggle switch, S1. A 1000-ohm fixed resistor is inserted in a leg of each toggle switch, which serves to limit the maximum line current to approximately 100 milliamperes.

The local current is required only when using radio adapters that do not supply their own line current. When the line current is supplied by the radio adapter, the "local-external current" toggle switch is placed in the EXT position.

FREQUENCY-SHIFT KEYER KY-75/SRT

The KY-75/SRT frequency-shift keyer (fig. 10-6) is a calibrated frequency-shift exciter in which signaling is accomplished by shifting a constant amplitude carrier between two fixed frequencies. The two frequencies represent the marking and spacing conditions of the telegraph code.

The keyer functions so that the current impulses from the teletypewriter keyboard produce a marking signal. This action causes the transmitter to emit a frequency above its mean frequency. Conversely, the opening of the contacts on the keyboard produces a spacing signal. The spacing signal, or open keyboard, causes the transmitter to emit a frequency below its normal assigned frequency.

The primary purpose of the frequency-shift keyer is to replace the conventional exciter of a c-w transmitter with a source of r-f excitation that can be shifted (in frequency) a small amount upward and downward to produce r-f telegraph code. The unit responds to both neutral (d-c) and polar input signals. Neutral signals cause a current flow in the line during the marking condition and no current flow during the spacing condition. Polar signals allow current in one direction through the line for mark and in the opposite direction for space. The keyer is used principally for comparatively long distance communications in the high-frequency range.

The keyer is composed of two subunits, which are the modulator-power supply and the amplifier-oscillator. The modulator-power supply (lower panel) incorporates the power supply and all modulator circuits up to but not including the reactance tube (discussed later). The amplifier-oscillator section (upper panel) comprises the crystal oven, reactance tube, 200-kc oscillator, and all r-f circuits and controls. All frequently used operating controls are located on the front panel. Semioperating controls are mounted on a subpanel behind the front panel.

Transmitters such as the AN/SRT-14, 15, and 16 (discussed in chapters 5, 6, and 7) do not require the use of an external frequency-shift keyer to produce telegraphic frequency-shifted signals. This is true since these transmitters receive the mark and space signals at an internal frequency-shift oscillator, which, in turn, shifts the output frequency of the transmitter accordingly.

BLOCK DIAGRAM

The input radio frequency to the frequency-shift keyer may be derived from an external oscillator, such as the master oscillator of the transmitter, or from the self-contained crystal oscillator (fig. 10-7). The over-all output frequency range of the keyer is 1.0 to 6.7 megacycles. Since the r-f input is mixed with a 200-kc signal from the frequency modulated oscillator to obtain the sum frequency, the input frequency must range from 0.8 to 6.5 megacycles.

The frequency of the 200-kc oscillator is frequency modulated by the balanced reactance modulator. The reactance modulator varies the amount of reactance across the oscillator tuned circuit, thus causing the oscillator frequency to increase or decrease a small amount in response to mark and space signals.

The magnitude of the frequency shift is adjustable over a range from 0 to ±500 cps or 1000 cps total shift. The transmitter frequency may be adjusted to any value from 0 to 500 cps higher than the assigned frequency for the mark signal and to the same number of cycles lower than the assigned frequency for space. The

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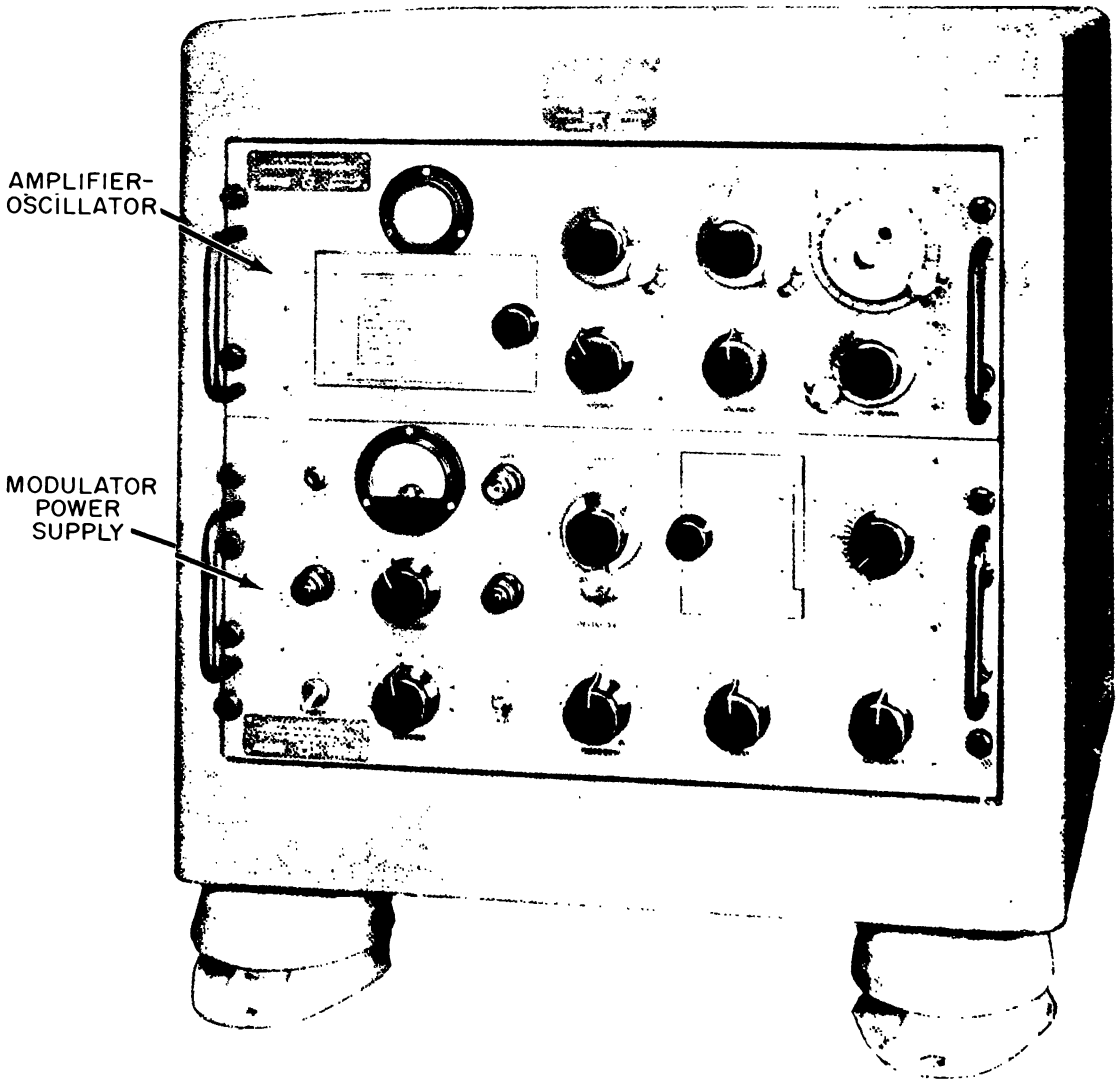

Figure 10-6. —Frequency-shift keyer KY-75/SRT.

equipment is capable of being keyed up to 240 dot-cycles (transitions from mark to space and vice versa) per second.

The radio frequency output of the crystal oscillator and the frequency modulated output of the 200-kc oscillator are combined in the balance mixer circuit. The frequency of the crystal oscillator is balanced out. Therefore, only the sum and difference frequencies resulting from the mixing of the two oscillator frequencies are present in the output of the balanced mixers. The circuits of the balanced mixers are tuned to the sum frequency.

The buffer amplifier serves to isolate or permit light loading at the balanced mixer output circuit, thus providing greater discrimination against unwanted modulation components. A ganged tuning control simultaneously tunes the output of the crystal oscillator, balanced mixers, and the buffer amplifier. The power amplifier supplies an output of 6 watts into a 75-ohm noninductive resistive load.

The test-operate and calibrate circuit comprises a five-position switch utilized to select the circuit arrangement required for frequency-shift or facsimile operation. It also selects the arrangement required to perform the alignment adjustments for carrier, mark, and space.

Two diode pulse limiters are employed to furnish a modulating wave effectively free from

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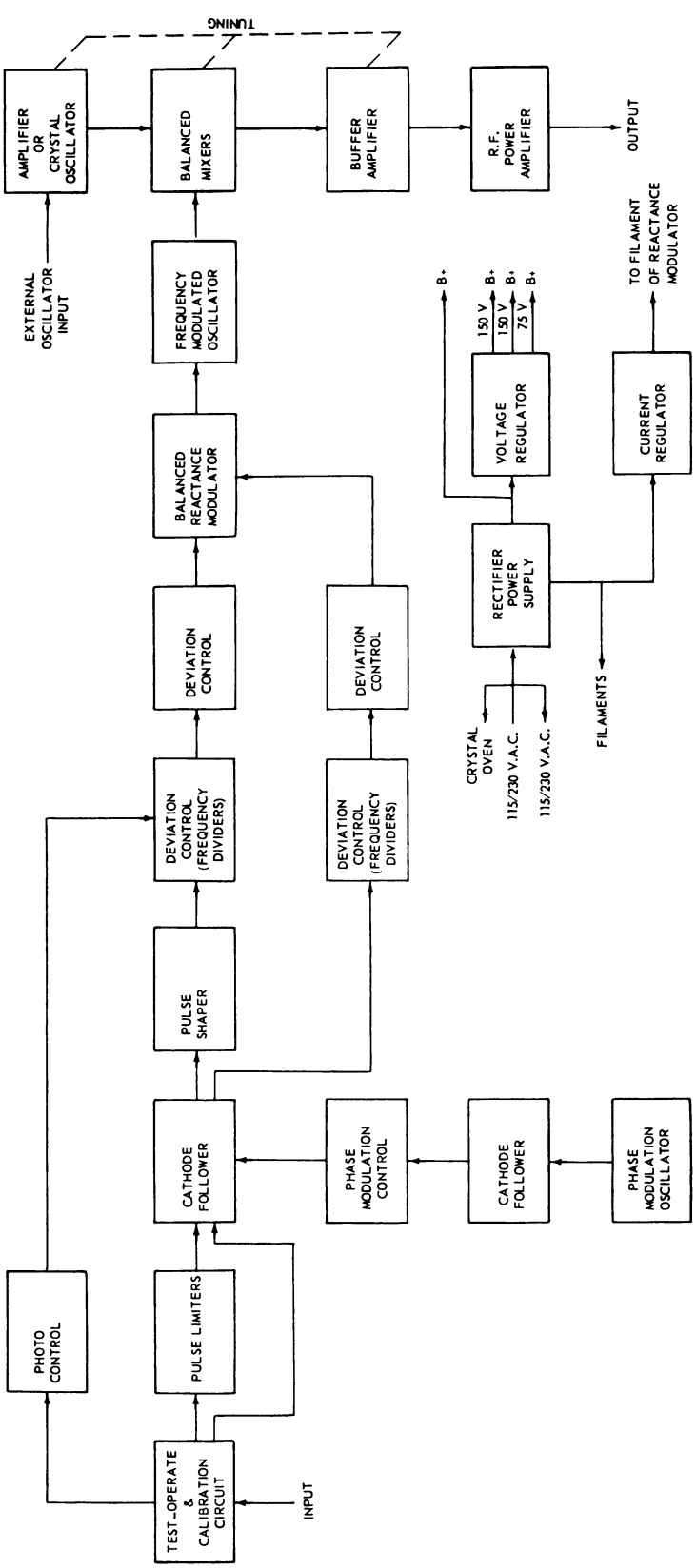

Figure 10-7. —Frequency-shift keyer block diagram.

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amplitude and wave front variations of the incoming telegraph signal. The output of the second limiter is coupled to a cathode follower, which acts as an impedance transforming device to match the low impedance pulsing circuits to the relatively high impedance calibration circuits.

The phase-modulation oscillator supplies a 200-cycle phase-shift signal, which provides a means of reducing the effects of selective fading during adverse operating conditions. However, phase modulation, which causes a slight shift in frequency, is normally not used if the total frequency-shift between mark and space signals is less than 600 cps or, if the dotting speed is faster than 35 dot-cycles per second. The phase-modulation control provides a means of adjusting the shift in phase from zero to one radian (57.3°).

The frequency-shift signal transitions are wave shaped to restrict sideband radiation by means of four low-pass filters incorporated in the pulse shaper. If the square waves from the limiting circuit are rounded before being passed to the reactance modulator, there are fewer sidebands generated because of the partial elimination of the harmonic content of the square wave. The reduction of sidebands permits the assignment of closer radio channel spacing for telegraph operation with the keyer.

During high-frequency transmission when the r-f carrier frequency is multiplied by some factor in the transmitter, the amount of deviation of frequency shift is multiplied simultaneously by the same factor. It therefore becomes necessary to reduce the amount of deviation by a factor equal to the multiplication factor of the transmitter. This action is accomplished by the deviation multiplier, which provides a means of dividing the frequency deviation by a factor of 1, 2, 3, 4, 6, 8, 9, or 12. Thus, the transmitter output is held to a certain preset deviation regardless of the frequency multiplication employed in the transmitter.

The deviation control circuits act to keep the wave-shaping termination at a fixed value. The calibrated dial of the deviation control reads the actual amount of deviation realized at the output of the transmitter.

When the test-operate switch is placed in the PHOTO position, the equipment is in readiness for facsimile operation. Photo facsimile operation is similar to frequency-shift key (FSK) operation, except that during photo operation, the limiting and wave-shaping circuits are not utilized. The photo control circuit provides a means of adjusting the photo input voltages.

Throughout this chapter, emphasis is placed on the circuit operation during frequency-shift keying. Facsimile operation will be discussed only when such discussion hinges upon, or is directly related to, the action during frequency-shift operation.

The power supply contained in the modulator-power supply is designed to operate from a 115- or 230-volt, a-c, single-phase, 60-cycle input source. The supply furnishes 270 volts d-c at 220 milliamperes, 4.35 amperes at 6.3 volts a-c, 0.6 amperes at 12.0 volts a-c, and 3 amperes at 5.0 volts a-c.

CIRCUIT OPERATION

Limiters

Two double-diode limiters, V1 and V2 (fig. 10-8) receive and shape the input d-c telegraph signals. The signals are applied from the key-line through the test-operate switch sections, and S1C and S1A (shown in the FSK position) and developed at the V1A cathode. Negative input signals cause V1A to conduct such that it reduces polar keyed signals to on-off signals. The portion of the signal developed across R18 is negative with respect to ground, and is removed completely by V2A.

Positive input signals are limited by V1B and V2B. The potential at the V1B cathode is maintained at approximately +20 volts. The potential is obtained from a voltage divider comprising R20 and R21, which are connected from ground to the regulated B supply. When the positive input signal exceeds +20 volts, V1B conducts, thus limiting the value of the voltage that can be applied to the V2B plate. The amount of the limiting action of V2B is determined by the positive voltage (approximately 12 volts obtained from the regulated B supply), which is maintained at the V2B cathode. The FSK mark adjust control, R23, is used to adjust the voltage change between mark and space signals.

Cathode Followers

The signal from V2B is applied through S1B to the control grid of the cathode follower, V3. During FSK operation, the mark voltage appearing at the pin 1 grid is approximately 8 volts. A fixed d-c potential of about 4 volts, which is equivalent to the average carrier voltage, is obtained from the regulated B supply (75 volts),

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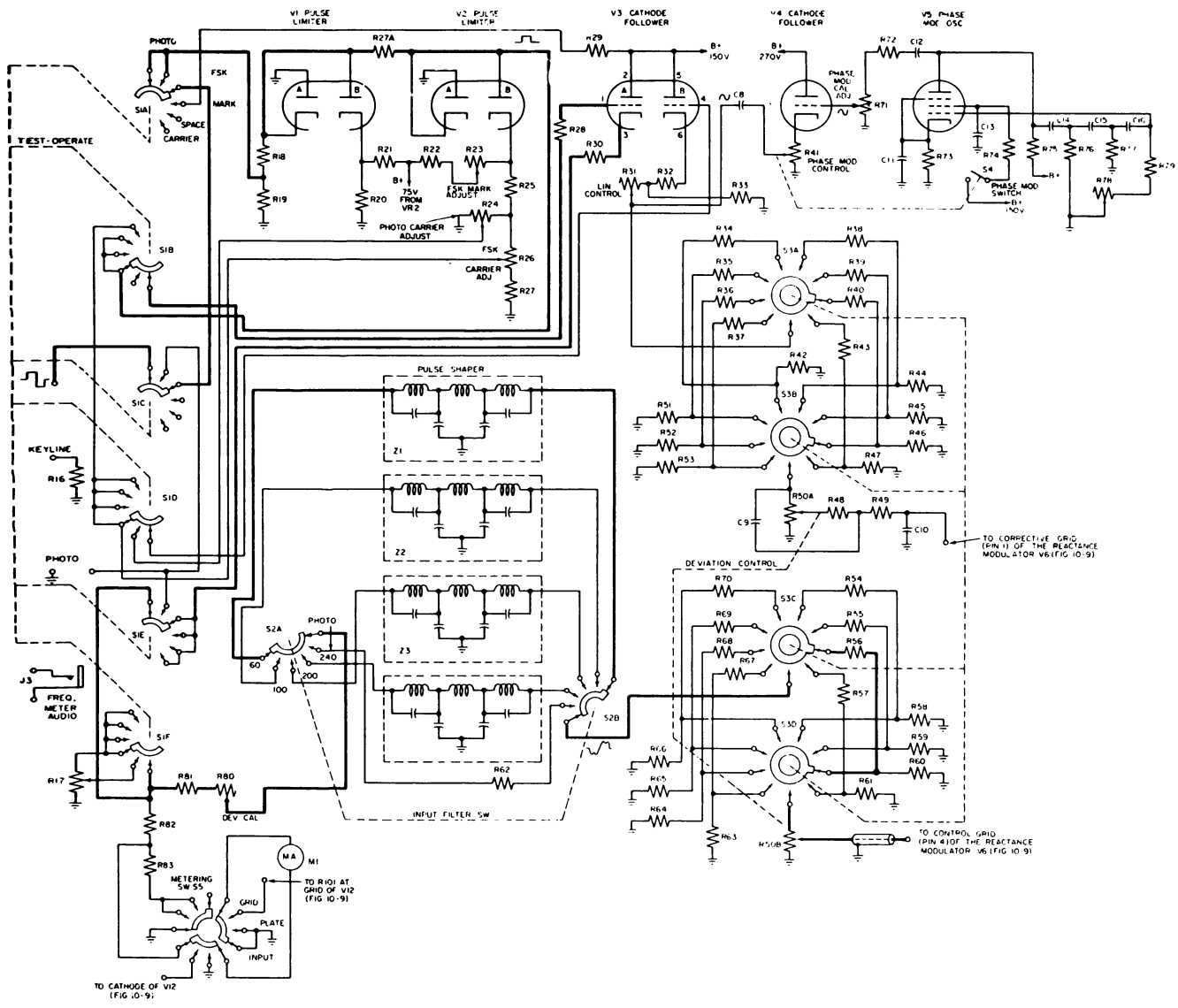

Figure 10-8. —Modulator of KY-75/SRT.

and applied to the corrective grid (pin 4) of V3. A portion of the d-c cathode voltage of V3B is applied directly to the reactance modulator (fig. 10-9). The reactance modulator, in turn, controls the frequency output of the keyer. FSK carrier adjust, R26, is used to correct the carrier frequency of the keyer as desired. R28 serves as a grid limiting resistor of V3A. When the test-operate switch is placed in the MARK position, R28 is connected in series with R27A and R18 through S1B and through S1A to R29 and the B supply.

The square pulses arriving at the V3A grid (fig. 10-8) are reproduced at the cathode and applied through R30, S1E, R81, and R80 to the input filter switch, S2A. Either of four filters may be selected by S2 to reduce the harmonic content of the square pulse. The filters are capable of 30 db attenuation to frequencies above the third harmonic of the keying speed. A fifth position of the switch disconnects the filter during photo facsimile transmission.

When the switch, S2, is set at position 60 (as indicated on the front panel in figure 10-6) the

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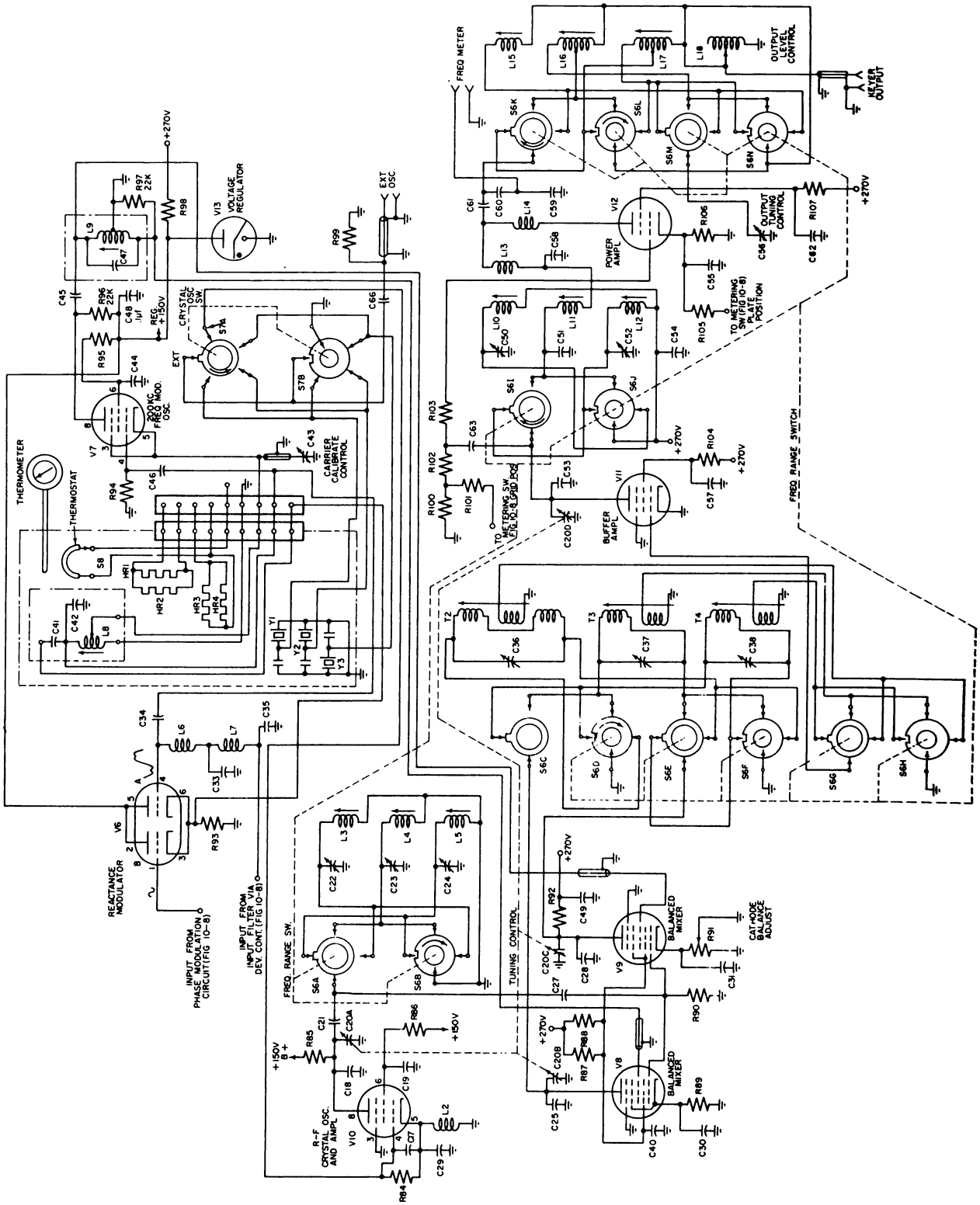

Figure 10-9. —Amplifier-oscillator of the KY-75/SRT keyer.

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filter provides a partial rounding of the square waves. The filter (pulse shaper) employs a low-pass pi network with parallel-tuned circuits connected in series with the input and output. The rounding of the wave is progressively less pronounced as the switch is turned to position 100 and then to 200. In position 240, practically unmodified square waves appear at the filter output.

Deviation Multipliers and Deviation Control

The deviation multiplier switch, S3, provides a method of dividing the frequency deviation of the keyer to allow for frequency multiplication in the transmitter. The deviation division is obtained through the use of a voltage divider network of series-shunt connected, noninductive resistors in conjunction with the multiplier switch.

The voltage divider network is so arranged that as the amount of deviation is changed, a small value of resistance, R46, is shunted across the deviation control, R50A, and a large resistance value, R40, is placed in series with the control. Various other selections of series and shunt resistances can be chosen by operating S3. In this manner, the wave-shaping filter termination of Z1 through Z4 is maintained at 5000 ohms.

The d-c potential existing across R50A is applied directly to the control grid of one-half of the reactance modulator (fig. 10-9). The potential developed across R50B (also positive with respect to ground as a result of the conduction of V3A) is applied to the second half of the reactance modulator. All sections of S3 are ganged together and changed simultaneously.

As the shunt resistance across R50A and R50B varies, the voltage applied to the reactance tube grids will also vary, which, in turn, shifts the output frequency of the keyer. As the amount of deviation is lowered, a larger resistance shunts the control, and a smaller resistance is placed in series. By reducing the shift in this manner, the mean carrier frequency is maintained at its center position between mark and space as the multiplication factor is altered. The calibrated dial of the deviation control reads the actual amount of deviation realized at the output of the transmitter.

Now assume that the keyer is used with a radio circuit in which a total frequency shift of 850 cycles between marking and spacing signals is desired. The carrier $\frac{+850}{2}$ equals the mark output frequency, and the carrier $\frac{-850}{2}$ equals the space frequency. The actual frequency shift necessary at the output of the keyer unit depends upon the frequency multiplication factor at the transmitter. If the transmitter mulitiplication factor is 3, the frequency-shift at the keyer output must be $\frac{850}{2x3}$ = 141 2/3 cycles.

The keyer is designed for a frequency shift that is adjustable from 0 to ±500 cycles for FSK operation. Any value of shift between 0 and ± 1000 cps with respect to the assigned frequency is available during photo facsimile transmission.

Photo Input Control

The limiting and wave-shaping circuits are not used during photo transmission. The photo signal is applied through S1C, S1E, and S1F (each in the PHOTO position) to the arm of R17 whose terminating impedance is about 600 ohms. R17 serves as the photo input control, and provides a means of adjusting the photo input voltage to a peak value of 5 volts before application to the frequency-shifting circuits.

The input voltage amplitude may be observed on the panel meter, M1, by setting the metering switch, S5, to INPUT. The grid position of S5 connects the meter in the control grid circuit of the power amplifier, V12 in figure 10-9, and in the plate position the meter is connected in the cathode circuit (measuring plate current) of the same tube.

The signal from S1F (fig. 10-8) is also coupled through R81 and R80 to the input filter switch, S2A. Since filtering is not required during transmission, the photo signal is coupled through S2A (in the PHOTO position) through R62, and through S2B (also in the PHOTO position) to the deviation control circuit. The signal is then coupled through R56 (or any of the other resistors selected by S3) and from the arm of R50B to the control grid (pin 4) of the reactance modulator, V6 (fig. 10-9).

Phase-Modulation Oscillator

The phase-modulation oscillator, V5 (fig. 10-8) produces a 200-cycle voltage, which may be used to produce a phase shift in the keyer output. The shift is useful to reduce the effects of selective fading during adverse operating conditions.

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V5 is switched into operation by the phase-modulation switch, S4, which applies the B supply voltage to the screen grid. The circuit of the oscillator is that of a conventional phase-shift oscillator.

The resistance-capacity, phase-shifting network is composed of capacitors C14, C15, and C16, and resistors R76, R77, R78 and R79 in series. The network is connected between the output and input circuits. The frequency control, R78, has a range of 40 cycles and is used to adjust the output frequency to 200 cycles.

The 200-cycle signal is developed across a voltage divider comprising C12, R72 and the phase-modulator calibration adjust, R71. The magnitude of the input signal to the cathode follower, V4, is determined by the setting of R71. The amplitude of the output signal from the cathode follower is controlled by the setting of R41. The signal coupled through C8 is superimposed on the voltage applied to the reactance modulator.

Reactance Modulator

During frequency-shift operation, a mark voltage of approximately 8 volts is applied to the signal grid (pin 1) of cathode follower, V3 (fig. 10-8). A fixed d-c voltage of approximately 4 volts, which is equivalent to the average carrier voltage, is connected from voltage regulator, VR1 (fig. 10-10) to the corrective grid (pin 4) of cathode follower V3 through the FSK carrier adjustment, R26. The corrective voltage is connected to the corrective grid (pin 1) of reactance modulator V6 (fig. 10-9), through one section of the corrective network consisting of C8 and the resistors selected by the deviation multiplier switch sections S3A and S3B. The oscillator signal voltage developed at the arm of the dual deviation control, R50A, is applied to pin 1 of the reactance modulator, V6 (fig. 10-9).

A correction circuit is necessary in the reactance modulator, V6, to maintain proportional voltage changes on both grids of V6 as the deviation is increased from zero. As the deviation is increased, the zero (space) signal frequency will decrease to the correct space value.

When the multiplier switch, S3, and the deviation control, R50, are varied, the voltage on the corrective grid (pin 1) of V6 is kept equal to the average carrier value of the signal grid (pin 4) of V6.

By controlling the circuit in this manner, the deviation can be changed without readjusting the carrier; whereas, in a noncorrected deviation circuit, the carrier will shift between the mark and space frequencies with changes in deviation, necessitating returning of the carrier.

The reactance modulator functions as a dual cathode follower whose input capacity at the pin 4 grid can be varied by changing the gain of the circuit between 0 and 1. This action changes the frequency of the 200-kc oscillator, V7, accordingly.

Section B of V6 controls the conduction of section A as a result of the common cathode resistor, R93. The input circuit of the B section of V6 is made to be capacitive by connecting C34 and C41 between the grid (pin 4) and the cathode (pin 6). The series combination of R93 and C41 is connected across the 200-kc oscillator tank circuit comprising L8 and C42. Under no signal condition the circuit constants are such that the oscillator frequency is held at 200 kc. As the gain of the cathode follower changes from 0 to 1, a varying current passes through C41, which varies its effective shunting capacitance. This causes the oscillator tank frequency to vary accordingly. Actually, the cathode follower gain is controlled by the bias on the grids of the A and B sections of V6.

200 KC Oscillator

The signal from the reactance modulator, V6, is coupled by C46 to the control grid of the 200-kc, frequency-modulation oscillator, V7. The circuit is that of an electron-coupled modified Hartley oscillator.

The main tuning tank of the oscillator comprising L8 and C42, along with the grid-to-cathode (input) capacitor, C41, is housed in a temperature-controlled oven. L8 is factory adjusted and normally will not require readjustment in the field. A thermostat and thermometer automatically control the oven temperature, which will be discussed later.

The carrier calibrate control, C43, is shunted across a portion of the tank inductor, L8, and provides a method of correcting the oscillator frequency to exactly 200 kc. The control compensates for minor changes in the crystals and/or the oscillator tube.

A regulated 150 volts is applied to the screen grid, which acts as the plate of the oscillator. One end of the pin 8 plate load resistor, R96, is held at r-f ground potential by C48. The 22-kilohm resistor shunts one-half of the output tank inductor, L9. The reactance of C45 at 200 kc

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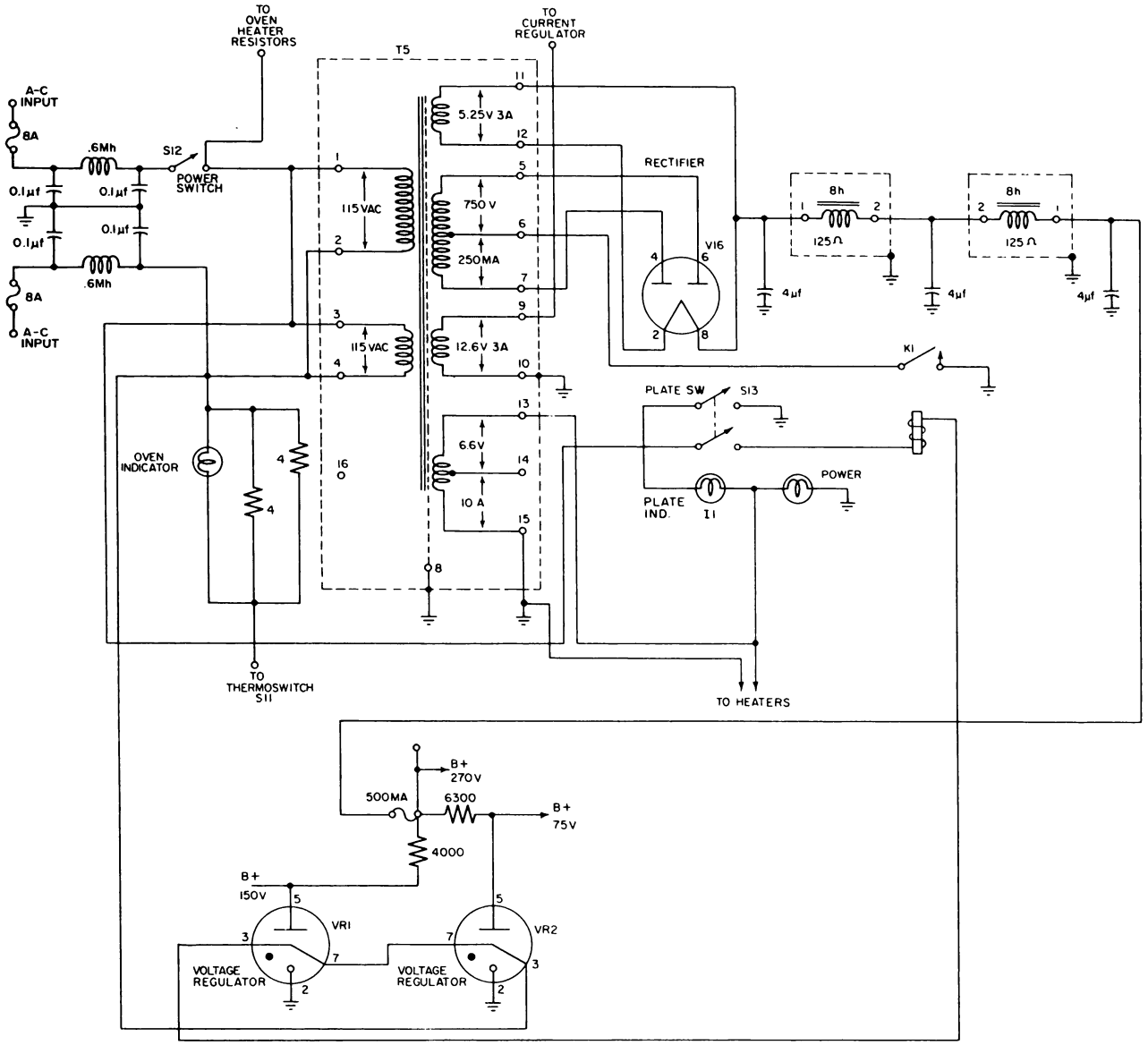

Figure 10-10. —Power supply of the KY-75/SRT keyer.

(approximately 79 ohms) is negligible. The other half of the tank inductor is shunted by R97. In this manner, equal amplitude output signals (180° out-of-phase) are applied from each end of L9 to the respective injector grids of the balanced mixers, V8 and V9.

Crystal Oscillator-Amplifier

The crystal oscillator-amplifier, V10 (fig. 10-9) is equipped with six crystal sockets located in the temperature controlled oven. Three are type HC-6/U crystal holders, and the remaining three are type HC-1/U. The sockets are arranged so that crystals cannot be plugged into adjacent holders at the same time.

The tuned circuits of the crystal oscillator, balanced mixers, buffer, and final amplifier are divided so that they cover the entire frequency range in three bands. The bands are divided to cover from 1 to 1.8 mc., 1.8 to 3.5 mc., and 3.5 to 6.7 mc, respectively. Either one of the bands may be selected by the frequency range switch, S6.

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Either of three crystals or an external oscillator may be selected by the crystal-oscillator switch, S7. When the switch is set to the EXT position (as shown) the external oscillator input will be applied through C66 and S7A to the control grid of V10. Thus, the stage acts as an amplifier when the switch is in this position. S7B grounds the unused crystals in each position of the switch.

Either of the remaining three positions of the switch, S7, will shunt one of the three crystals, Y1, Y2, or Y3, across the V10 input circuit. The circuit can therefore be described as a Colpitts crystal-controlled, electron-coupled oscillator.

A parallel-tuned tank made up of L2 and C29 is placed in the V10 cathode circuit. The circuit is provided for feedback purposes only. L2 completes the d-c cathode-to-ground path, and, in addition, forms a resonant circuit with C29 whose resonant frequency is below that of the chosen crystal. The circuit is therefore capacitive, and allows sufficient feedback to sustain oscillations. This circuit arrangement is common to Colpitts oscillators used at relatively high frequencies.

The screen grid (pin 6) of V10 serves as the plate of the oscillator. The virtual cathode existing between the screen and suppressor grids passes the electrons (at the oscillator frequency) to the plate (pin 8). The output of the oscillator is coupled through C21 and C27, and is developed at the control grids of the balanced mixers, V8 and V9, by R90. The oscillator is tuned by C20A, which is ganged to the tuning controls of the balanced mixers and the buffer amplifier.

Balanced Mixers

The balanced mixers employ two pentagrid tubes, V8 and V9, which receive the 200-kc, frequency-modulated signal from V7 at their respective injector grids, and the crystal oscillator V10 output at the control grids. The two signals are heterodyned in the mixers to produce the frequency of the final carrier output. The balanced mixer output is subsequently acted upon by the buffer, V11, and the final amplifier, V12.

The 200-kc, frequency-modulation oscillator input at the grids of the balanced mixers, V8 and V9, are 180° out-of-phase as a result of the center tapped output coil, L9, in the V7 plate circuit. The modulation components in the balanced mixer output is likewise 180° out-of-phase with each other and are fed to opposite ends of the tank, T4. The signals fed to the tank act in series aiding across the output transformer. The r-f crystal oscillator output components of V8 and V9 are in phase with each other, and thus cancel across the output transformer.

The crystal oscillator frequency is balanced out by the cathode balancing adjustment, R91, which is used to equalize the gain of the crystal oscillator frequency in the two tubes. The plate circuits of the mixers are tuned by sections B and C of the tuning capacitor, C20. The T2, T3, and T4 primaries with trimmers C36, C37, and C38 are selected respectively by the frequency range switch, S6, to complete the mixer tuned circuits. When the cathode balancing adjust, R91, is properly set, tuning of the plate circuit will produce a minimum amount of grid drive at the crystal oscillator frequency. Sections S6D, S6F, and S6H of the frequency range switch grounds the unused plate components of V8 and V9.

The sum and difference frequencies resulting from the heterodyning action are present in the output of the balanced mixers. The plate tank circuit is tuned to the sum frequency.

Buffer Amplifier

The output of the balanced mixers is applied through the selected section of S6 to the control grid of the buffer amplifier, V11. The buffer stage minimizes the loading on the mixer output circuit, and, together with the balanced mixer tuned plate circuit, provides greater discrimination against unwanted modulation components.

The plate tank of V11 comprises inductors L10, L11, and L12; trimmers C50, C51, and C52; and section D of the tuning capacitor, C20. The desired tuning inductor and trimmer capacitor are selected by sections S6I and S6J of the frequency range switch. The buffer output is coupled by C63 through parasitic suppressor R103 to the control grid of the power amplifier, V12.

Power Amplifier

The power amplifier, V12, is a class "C" operated stage, which boosts the signal amplitude to the required output level of the keyer. The plate circuit of V12 contains a series peaking coil, L14, which couples the amplifier output through C61 to the frequency range switch. Section S6K of this switch selects either of the output inductors, L15, L16, or L17. S6M shunts the output tuning capacitor, C56,

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across the chosen inductors. S6L and S6N short the unused inductors.

The output of the selected plate tank is coupled to the output level control, L18, which matches the keyer output to the impedance of the transmission line. The keyer output terminal is a coaxial jack from which the frequency-shifted keyer signal may be connected to the transmitter.

A panel meter (fig. 10-6) may be connected in the grid or plate circuit of the power amplifier through the use of the metering switch, S5 (fig. 10-8). Thus, visual indication of these currents in the final amplifier circuit can be obtained for tuning purposes.

A capacitor voltage divider comprising C59 and C60 is connected between the plate and ground of V12 (fig. 10-9). The signal voltage at the junction of the two capacitors is fed to frequency meter jack for application to frequency measuring equipment.

Crystal Oven

The crystal oven is provided to assure minimum change in the value of the frequency determining components of the crystal and 200-kc oscillators. The oven is composed of four plaque resistors, HR-1, HR-2, HR-3, and HR-4, which operate from 115 or 230 volts obtained from the keyer power supply. The four heaters are insulated from the outer walls by strips of asbestos. The oven contains the crystal holders, the 200-kc oscillator tank circuit, and the cathode-to-grid capacitor, C41, of the reactance modulator.

The temperature of the oven is controlled by a thermoswitch, S8. The S8 contacts are closed, keeping the lamp lighted and the heater operating until the temperature of the oven rises to 70° C. The contacts open and close intermittently to maintain the temperature at this level. A -50- to +100-degree centigrade thermometer and an oven indicator lamp (not shown) are utilized as oven temperature indicators. The arrangement of oven heaters provides a constant temperature but not an extremely rapid warmup.

Power Supply

The power supply of the KY-75/SRT frequency-shift keyer (fig. 10-10) is a conventional, full-wave rectifier employing a double-pi filter. A-c power input is applied by the main power switch, S12. The power supply is wired for 115-volt operation, but, with minor changes, it can be adapted for a 230-volt, a-c input.

In 115-volt operation, the two windings of the T5 primary, and the oven heater resistors (fig. 10-9) are connected in parallel across the a-c input. During 230-volt operation, the two windings and the heater resistors are connected in series.

A relay, K1, is provided in the B- circuit of the power transformer. When the plate switch, S13, is closed, the a-c power circuit is completed through the plate indicator lamp, T1, and through the K1 solenoid. K1 becomes energized, closing its contacts, and thereby completing the B-circuit.

Two voltage regulator tubes, VR1 and VR2, are employed to increase the stability of the voltage applied to the keyer circuits. The circuit supplying the power to the K1 solenoid is completed through pins 3 and 7 of the regulator tubes. If either of the tubes is removed from the power supply, plate and screen power cannot be applied to the keyer stages.

TRANSMITTER-TELETYPEWRITER CONTROL UNIT C-1004A/SG

The Transmitter-Teletypewriter Control Unit C-1004A/SG (fig. 10-11) contains the components and circuitry necessary for controlling a teletypewriter radio circuit from a remote position. The unit provides the transmitter power-on switch, the power-on indicator lamp, carrier-on indicator lamp, and a three-position rotary selector switch.

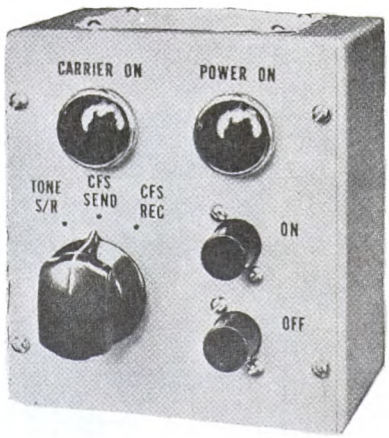

Figure 10-11. —Front panel of the transmitter-teletypewriter control unit, C-1004/SG.

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The circuit diagram of the control unit is shown in figure 10-12. The power switch, S1, permits power to be supplied to the remote transmitter through terminals 1, 2, and 3 of TB2. The power indicator lamp is lighted when the transmitter is on. S1 is a momentary contact switch.

S2 completes the circuit from a send-receive teletypewriter to either a frequency-shift keyer (CFS SEND), a frequency-shift converter or comparator (CFS REC), or a tone terminal on a send-receive basis (TONE S/R). The positions of the switch are shown in this order in figure 10-12, A, B, and C. The input d-c telegraph signals are applied at the teletypewriter signal line input terminals, A and B, of TB1.

When S2 is placed in the CFS SEND position (fig. 10-12, A) the teletype loop passes through terminal B of TB1, and through S2B to the frequency-shift keyer via terminal F. From the keyer, the loop continues through terminal E of TB1, through S2B and terminal A of TB1 to complete the loop circuit. Note that the frequency-shift converter (not in use) is shorted through S2C, and the tone terminal equipment (also not in use) is shorted through S2A. Thus, complete isolation of the unused terminals is ascertained.

Likewise, with S2 placed in the CFS SEND position, the transmitter carrier is turned on via S2C and terminals 5 and 6 of TB2. The circuit for the carrier-on indicator lamp is continued from one side of the line input voltage through S2B and terminal 4 of TB2 to the transmitter. The circuit is completed through the transmitter and terminal 2 of TB2 to the other side of the a-c line.

Figure 10-12, B and C, show the position of the S2 contacts for the CFS REC and TONE S/R positions, respectively.

When the switch is set to the CFS REC position (fig. 10-12, B) the carrier-on indicator lamp and the carrier are off. The teletypewriter is connected to the frequency-shift converter circuit by S2B, and the tone and frequency-shift keyer terminals are shorted by S2A and S2C, respectively.

In the TONE S/R position of S2 (fig. 10-12, C), the carrier-on indicator lamp and the transmitter remain deenergized. The teletypwriter is connected to the tone terminal loop through S2B. The frequency-shift converter and frequency-shift keyer are shorted by S2C.

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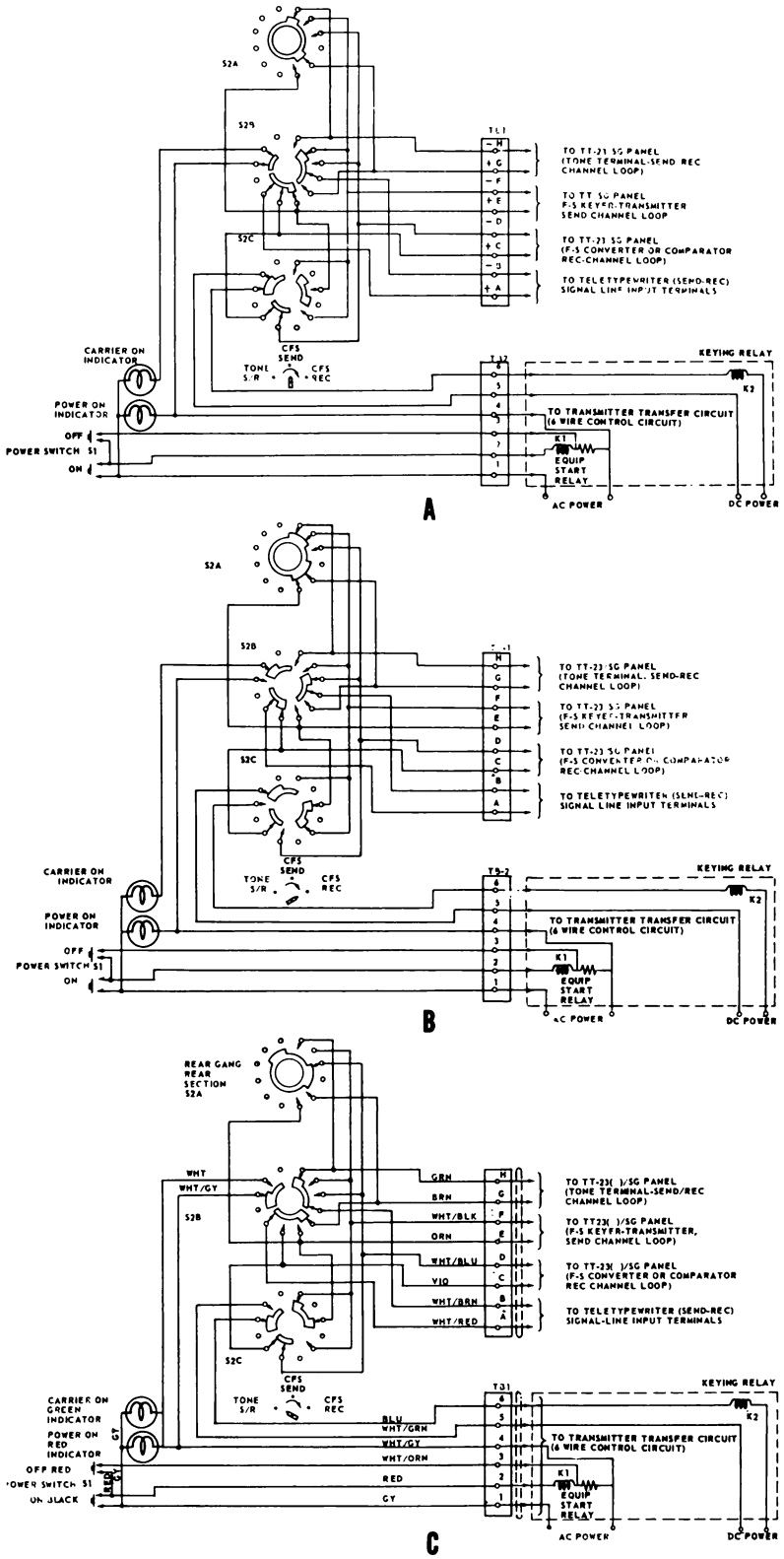

Figure 10-12. —Control unit circuit diagram.

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QUIZ

1. A radio teletype system is a series-connected telegraphic communication network operating between two or more points, and involving the transmission and reception of what type signals?
2. All components in the transmitting or receiving teletype system are series-connected, depending on the setting of what switch (fig. 10-1, A)?
3. When the page printer keyboard produces a marking signal, the output from the KY-75/SRT causes the transmitter to emit what frequency?
4. The frequency-shift keyer transmits a type F1 frequency-shift signal in what frequency range?
5. The control unit (fig. 10-13) provides what function?
6. The converter-comparator (fig. 10-15) is designed primarily for use during what type reception?
7. Rectification of the output voltage is accomplished by what components (fig. 10-3)?
8. To counteract the tendency of an output voltage rise, (fig. 10-3) the series impedance of coil 1-2 and coil 3-4 must be made to move in what direction?
9. Operating voltage for V2 (fig. 10-3) is obtained from a separate power supply, which employs V1 as the rectifier tube and what terminals of the power transformer?
10. The Teletype Panel TT-23/SG (fig. 10-5) is intended for general shipboard use for providing what function?
11. The insertion of a plug in the "set" jack opens the circuit from what unit (fig. 10-6)?
12. (a) The KY-75/SRT frequency-shift keyer (fig. 10-7) is a calibrated frequency-shift exciter in which signaling is accomplished by what method?
    (b) The keyer functions so that the current impulses from the teletypewriter keyboard produce what type signal?
13. Transmitters such as the AN/SRT-14, 15, and 16 do not require the use of a frequency-shift keyer since these transmitters receive the mark and space signals at what type circuit?
14. The frequency of the 200-kc oscillator (fig. 10-8) is frequency modulated by a signal from what stage?
15. The radio frequency output of the crystal oscillator and the frequency modulated output of the 200-kc oscillator (fig. 10-8) are combined in what circuit?
16. The power amplifier (fig. 10-8) supplies an output of what value?
17. The frequency-shift signal transitions (fig. 10-8) are wave shaped to restrict sideband radiation by means of four low-pass filters incorporated in what circuit?
18. During FSK operation, the mark voltage appearing at the pin 1 grid of V3 (fig. 10-9) is approximately what value?
19. The deviation multiplier switch, S3 (fig. 10-9) provides a method of dividing the frequency deviation of the keyer to allow for what function?
20. The d-c potential existing across R50A (fig. 10-9) is applied directly to the control grid of what stage?
21. The keyer, KY-75/SRT, is designed for a frequency-shift that is adjustable between what limits for FSK operation?
22. The reactance modulator, V6 (fig. 10-11) functions as a dual cathode follower whose input capacity at the pin 4 grid can be varied by changing the gain of the circuit between what limits?
23. (a) The tuned circuits of the crystal oscillator, balanced mixers, buffer, and final amplifier (fig. 10-11) are divided so that they cover the entire frequency range in how many bands?
    (b) What are the frequency limits of each band?
24. The crystal oscillator frequency of V8 and V9 (fig. 10-11) is balanced out by what control?
25. In the TONE S/R positon of S2 (fig. 10-12, C) the teletypewriter is connected to the tone terminal loop through what switch section?

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CHAPTER 11

RADIO TELETYPE SYSTEM, PART II

FREQUENCY-SHIFT CONVERTER-COMPARATOR AN/URA-8B

The frequency-shift converter-comparator, AN/URA-8B (fig. 11-1) is designed primarily for use with two standard Navy receivers in dual diversity reception of frequency-shift transmissions. The unit converts audio frequency-shift signals into pulses that are used to key the d-c loop circuit, thus causing the automatic teletype printers to reproduce the transmitted message. A keyed tone output signal is also produced for feeding the intelligence to remotely located telegraph or teletype terminal equipment over wire lines or radio links.

The radio receiver changes the received teletype modulated r-f carrier input to audio tones by means of the beat frequency oscillator. From this process, the r-f carrier shift becomes an audio frequency shift, which corresponds to the original shift produced at the distant frequency-shift keyer or transmitter output.

The converter-comparator, AN/URA-8B, is designed to operate from the audio output of two Navy-type RBA, RBB, RBC, AN/SRR-11, 12, 13, 13A, or any similar standard Navy radio receivers. The equipment will operate with the receivers in either space diversity or frequency diversity on carriers within the frequency ranges of the receivers employed.

The AN/URA-8B comprises two frequency-shift converters and one comparator. In diversity reception, the output of each receiver is connected to one of the two converters, and the d-c signals from the discriminator circuits of the two converters are fed to the comparator. In the comparator, the two signals are examined in a circuit, which automatically selects the better mark and the better space pulse for each character. In this manner, optimum characters are obtained from diversity reception.

The frequency-shift converters may be used separately for single receiver receptions of the frequency shift signals. Each converter has its own output circuits for keying the teletype d-c loop and for providing a keyed audio tone. When the converters are on single-receiver operation, the comparator may be associated with one of them to provide an additional set of output circuits, if desired.

The frequency shift employed in the converters may be as little as 10 cps or as much as 1000 cps frequency separation between mark and space signals. This scope of frequency shifts is divided into two ranges called, narrow-shift (10 to 200 cps) and wide shift (200 to 1000 cps).

BLOCK DIAGRAM

An input filter at the input of the frequency shift converter (fig. 11-2) eliminates the possibility of false keying by spontaneous noise signals outside the frequency-shift range of the converter. Two separate filters are used. The filters comprise a high-pass and a bandpass filter, respectively. The frequency response characteristics for the two filters are given in Chapter 13 of *ET-3*, NavPers 10188-A.

The limiter amplifier is designed to apply a constant voltage to the input circuit of the discriminator while variations at its input may swing from 60 μv to 60 mv. Amplitude limiting of both positive and negative peaks is accomplished in the circuit. The limiter output is coupled to the input of the discriminator circuit.

The discriminator circuit is essentially a double-slope detector consisting of three sections: The discriminator input network, the discriminator rectifier, and the discriminator buffer amplifier. The purpose of the discriminator circuit is to convert frequency-shift changes into corresponding d-c changes.

Two separate discriminator input networks are employed. One provides for narrow, frequency-shift operation, and the other passes the wide frequency-shift signals. The output of

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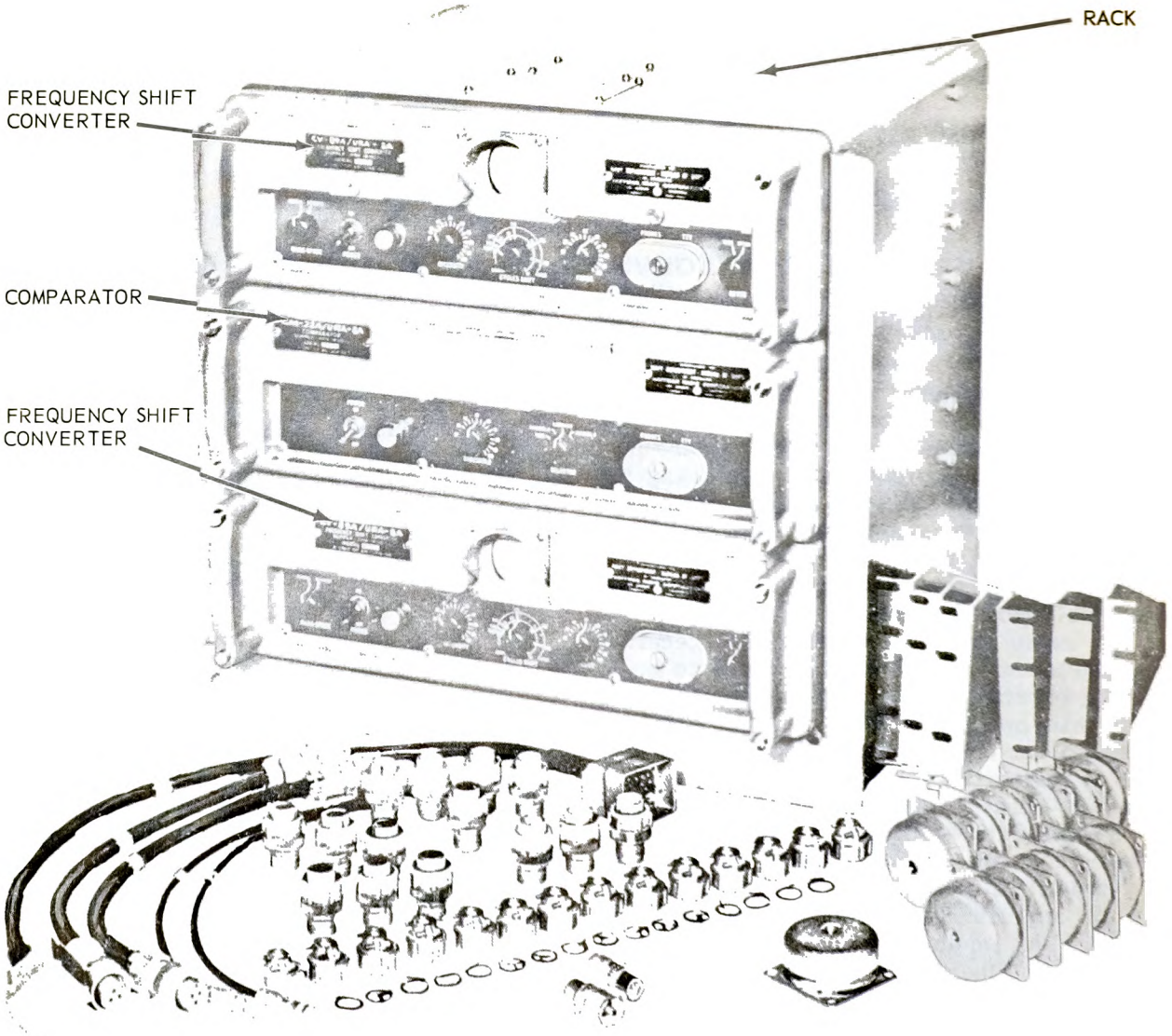

Figure 11-1. –Front panel and controls of the AN/URA-8B.

the two discriminator networks is applied to the buffer amplifier.

The fast-slow filter consists of two separate low-pass filters corresponding to a slow-speed filter and a high-speed filter, respectively. These circuits attenuate spurious signals above the frequency of the pulse rate of the circuit, preventing faulty keying action due to noise or harmonics.

The axis restorer circuit is provided to produce the optimum signal output when the received signal is weighted heavily on one side or the other, either mark or space. It also maintains the optimum axis, or bias, for keying with the weakest portion of a fading signal. The weighting of the signal may be due to the relative mark and space in each character, to selective fading conditions, or to improper tuning of the receiver. The axis restorer contains a system that "locks-up" the teletype circuit (closing the loop circuit) whenever there is a prolonged mark, space, or no-signal condition.

The oscillator-keyer subunit provides the necessary circuits for producing the tone output

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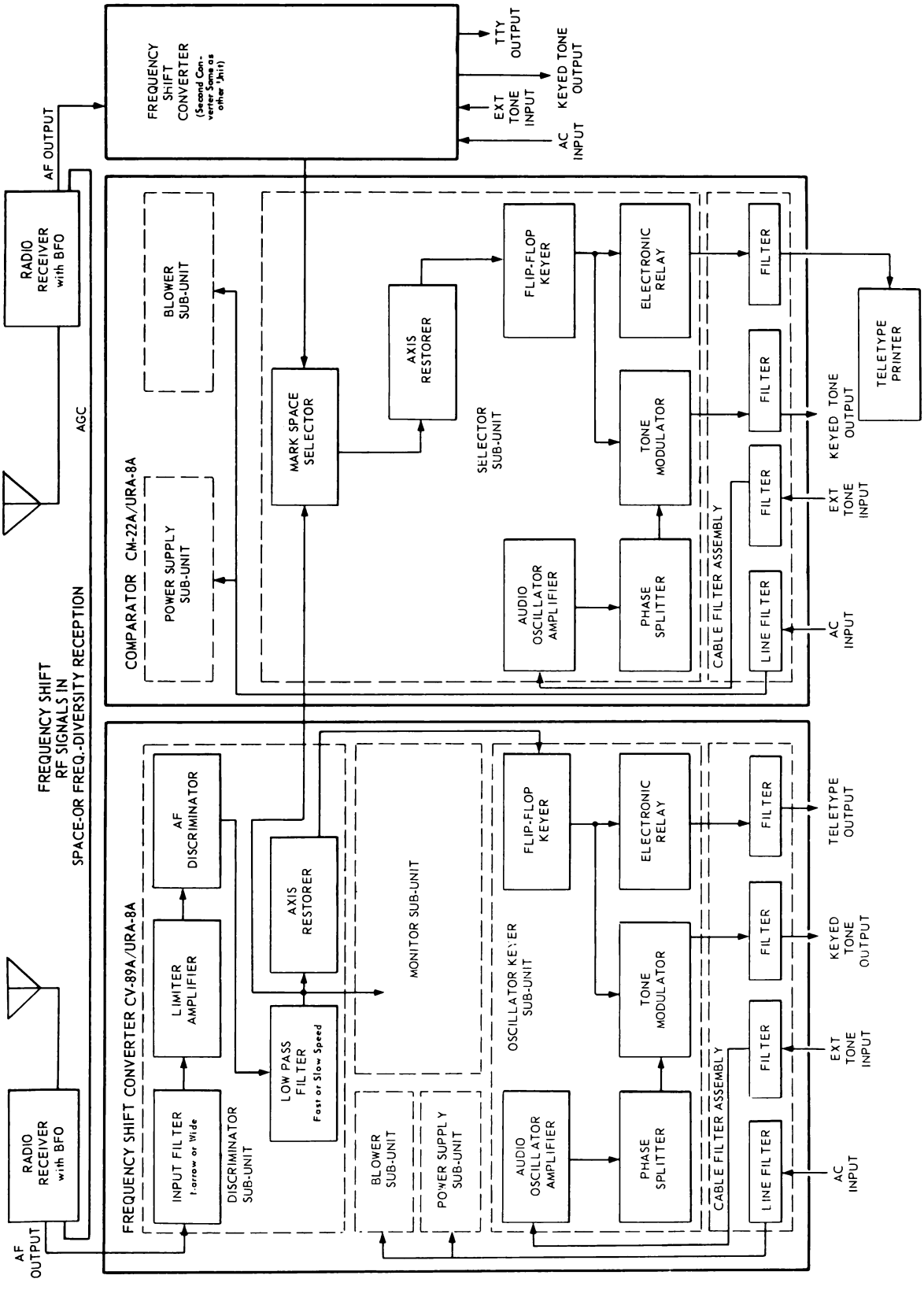

Figure 11-2. —Frequency shift-converter-comparator, AN/URA-8B, block diagram.

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to a tone receptacle and headphones, and for introducing the d-c pulses from the discriminator circuit to the teletypewriter printer. The audio oscillator-amplifier produces frequencies tunable to 595, 765, 935, 1105, 1275, 1445, 1615, and 1785 cps.

A phase splitter is included to supply the push-pull input circuit of the tone modulator. The latter is a conventional push-pull amplifier whose output is keyed from "tone-on" to "tone-off" by a signal from the flip-flop keyer circuit.

In addition to supplying a signal to the tone modulator, the flip-flop keyer circuit introduces the mark and space signals from the discriminator circuit to the electronic relay. Signals from the electronic relay circuit open and close the d-c loop of the teletype printer.

The power supply circuit of the frequency-shift converter is a conventional full-wave rectifier with a single pi filter. The B supply voltage supplied by the circuit is approximately 250 volts d-c. A negative bias of 38 volts is provided through the connection of a resistor and filter capacitor between the center tap of the high-voltage secondary and ground. The circuit of the power supply is not shown in this chapter. The complete schematic diagram of the power source can be found in *Frequency Shift Converter Comparator, Group AN/URA-8B*, NavShips 91490.

During diversity reception, the output signals from two frequency-shift converters are fed to the mark-space selector circuit of the comparator. This circuit automatically compares the two signals and selects the better mark pulse and better space pulse for each character of the coded message. The remaining circuits of the comparator perform the same function as described for the converter.

CONVERTER

Input Filter

Two input jacks, J1 and J2, respectively (fig. 11-3) receive the input from the receiver. The signal then passes through either of the two input audio filters, Z1 or Z2, corresponding to the band-pass and high-pass filters, respectively.

The bandpass filter, Z1, is used for the reception of wide frequency-shift signals and has an input impedance of 600 ohms. It should be recalled that the output from the AN/SRR-11, 12, and 13 receivers is fed through a 600-ohm balanced audio line. Direct connection of this line to either J1 or J2 will satisfy the input impedance characteristics of the converter.

The bandpass of Z1 for frequencies between 1450 and 3650 cps is within 6 db. Attenuation of frequencies below 1000 cps or above 4100 cps is greater than 40 db.

The high-pass filter, Z2, is used for the reception of narrow frequency-shift signals. The input impedance of Z2 is also 600 ohms. The characteristics of the filter are such that it is flat (+1 db) from 775 to 1400 cps, and down not less than 40 db at frequencies below 425 cps. Attenuation of frequencies above 1400 cps is provided in the audio amplifier circuits of most Navy receivers, and in the circuits of the converter. This action provides partial elimination of higher frequencies.

In the schematic diagram of figure 11-3 the narrow-wide switch, S1, is set for the reception of narrow frequency-shift signals, connecting Z2 as the input filter. Signals received at J1 are passed through the T1 primary by the circuit loop existing from terminal A of J1, through a section of the narrow-wide switch, S1A, through Z2 (pins 1 and 3), through another of the shorted sections of S1A, through T1 and another section of S1A, to terminal B of J1. The output of the input filter is therefore coupled to the limiter-amplifier by the matching transformer, T1.

Limiter-Amplifier

The limiter-amplifier consists of two resistance-coupled stages comprising the two sections of tube V1. R4 and R5, connected between ground and the B supply, provide a positive bias voltage at the V1A cathode, which places the tube operation at the midpoint on its characteristic curve. Strong signals overdrive V1A in the positive direction to cause limiting of the positive peaks, and strong negative signals are limited in the output circuit to the V1A cutoff.

Output signals at the plate of V1A are coupled by the voltage divider, C1-R7, to the V1B control grid. This stage operates as a conventional cathode-biased amplifier. The output signals of V1B are coupled by C2, through the B section of the narrow-wide switch, to the discriminator input network.

Discriminator Input Network

Two separate discriminator input networks (Z3 and Z4) are employed: One for narrow

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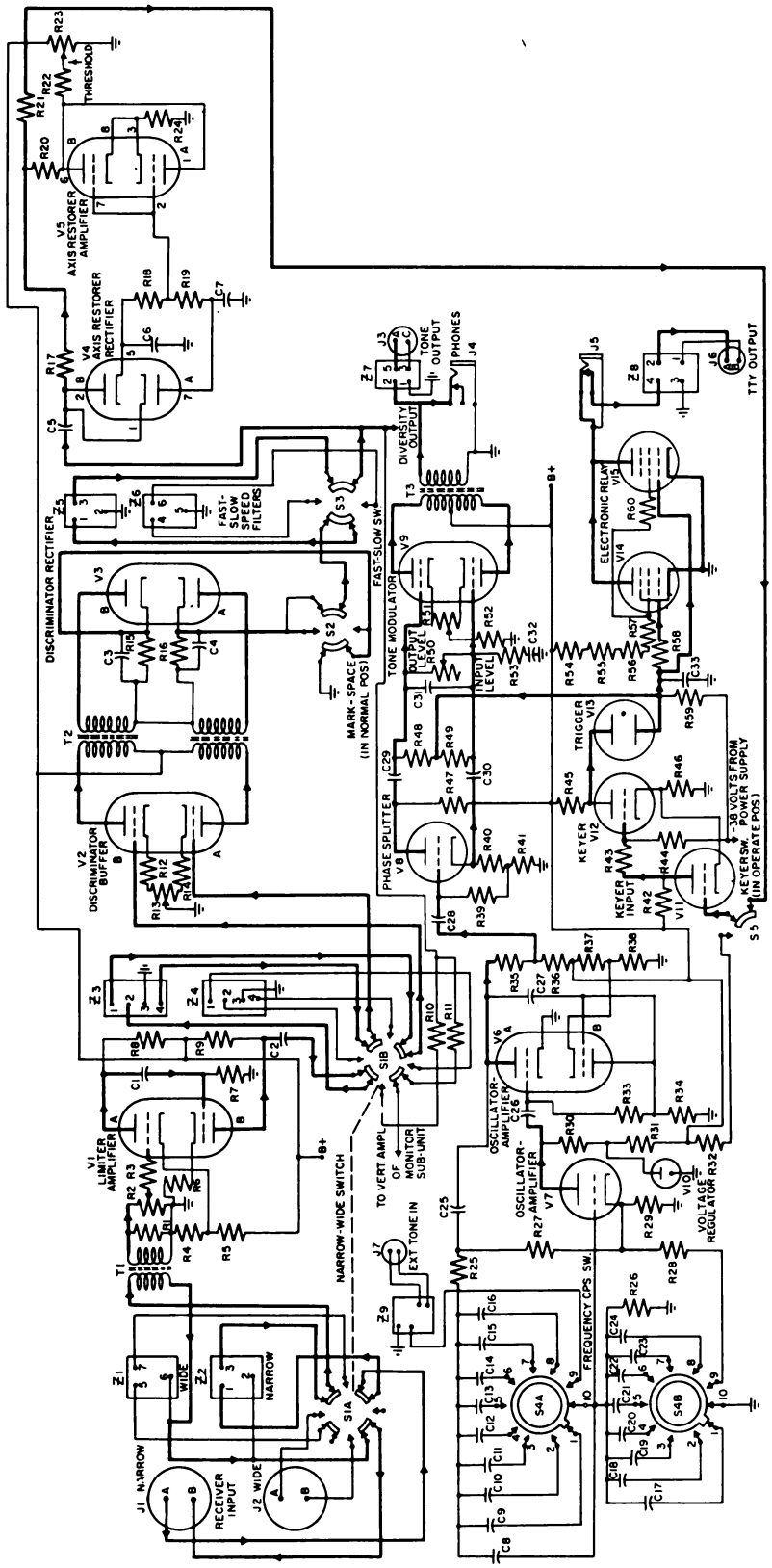

Figure 11-3. - Frequency-shift-converter, schematic diagram.

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frequency-shift and the other for wide frequency-shift operation. The narrow frequency-shift filter, Z3 (fig. 11-4) comprises two separate filters having a cross-over or center frequency of 1000 cps (± 1%). The filters are designed to cover the 10 to 200 cps frequency-shift range. Each filter comprises two series peaking tank circuits and two parallel tank circuits, one of which is tuned to 200 cps above the center frequency, and the other to 200 cps below the center frequency.

The frequency-shift network, Z4, is also composed of two filters and has a crossover frequency of 2550 cps (± 1%). These filters function for the 200 to 1000 cps frequency-shift range. Each of the two filters consists of a series (peaking) tank, and a damped parallel tank. One of the circuits is tuned to 750 cps higher than the center frequency, and the other is 750 cps lower. The output of the discriminator network filters is fed to the discriminator buffer amplifier (fig. 11-3) control grids through two sections of the narrow-wide switch, S1B.

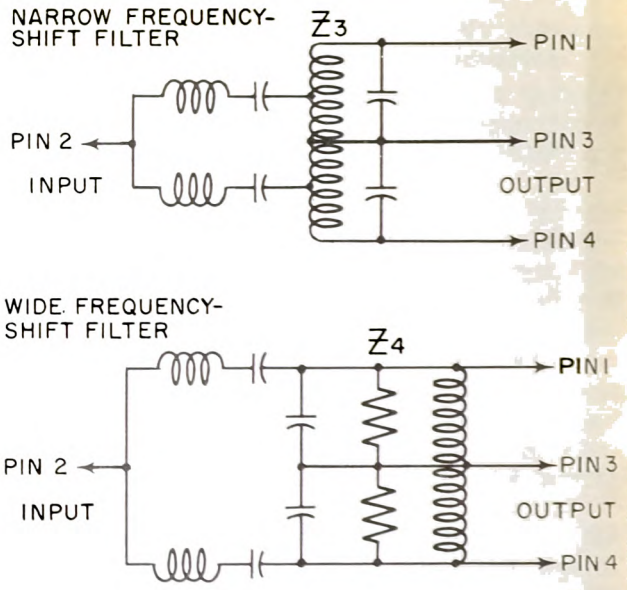

Figure 11-4. –Discriminator input network, simplified schematic.

Discriminator Buffer Amplifier

The discriminator buffer amplifier, V2A and B, consists of two separate low gain amplifiers. Bias voltage for both halves of V2 is obtained from R12, R13, and R14. R13 acts as the output balancing control for the two amplifiers and is adjusted for zero output at T2 when the center frequency is applied at the V2 input.

It should be recalled that the signal coupled to the converter input contains both high and low audio frequency components corresponding to mark and space signals, respectively. The grid of V2B is driven by the output from the lower frequency output terminal of the discriminator input network (pin 1 of Z3 or Z4). The grid of V2A is driven by the higher frequency component from pin 4 of Z3 or Z4. Thus, one stage of the buffer amplifier is amplifying the higher frequency output of the discriminator input network, and the other is amplifying the lower frequencies. The output of the separate amplifiers is coupled by separate sections of the output transformer to the discriminator rectifier circuit.

Discriminator Rectifier

The discriminator rectifier circuit comprises two diodes, V3A and B. V3B functions to rectify the low-frequency output of the buffer amplifier, and V3A rectifies the high-frequency output. Since the high and low frequencies are not rectified simultaneously, a single output connection at the V3A cathode couples the output signal through the mark-space switch, S2, to the fast-slow speed filter via S3. The signal output of V3 varies in magnitude and polarity corresponding to the differences in the audio input to the rectifier. S2 allows reversal of the mark-space conditions if the input signals are supplied from a maladjusted receiver.

The load resistors, R15 and R16, of the two rectifiers are of equal value to insure proportionate output signals. C3 and C4 bypass the higher frequencies (noise) around the load resistors.

Fast-Slow Speed Filter

The fast-slow filter consists of two separate low-pass filters, Z5 and Z6. These filters attenuate spurious signals above the frequency of the pulse rate of the circuit to prevent faulty keying action because of noise, harmonics, etc. The high-speed filter, Z6, is employed for keying speeds of 60 words per minute or more, and is flat (within 6 db) from 80 to 300 cps. The attenuation characteristics of Z6 at 500 cps and above are equal to or greater than 40 db.

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The slow-speed filter, Z5, is employed for keying speeds of 60 words per minute or less, and is flat (within 6 db) from 80 to 140 cps. The attenuation in this filter is up 40 db or more at frequencies of 240 cps and above. The output of the high- or slow-speed filter is delivered to the axis restorer and lock-up circuit (and to the comparator during diversity reception) through a section of the fast-slow switch, S3. The latter switch selects either Z5 or Z6, depending on the mode of operation.

Axis Restorer and Lock-Up circuit

The axis restorer circuit functions to improve the balance between the mark and space signal output. When no signal is received by the converter, either mark or space, the circuits of V4A and B and V5A and B receive a small positive voltage from the arm of R23. The positive potential existing at the R20-R21 junction is applied through the keyer switch, S5, to the control grid of the keyer tube, V11. This condition causes the keyer tube to simulate a lock-up condition, producing a marking signal at its output. The level of the voltage applied to the axis restorer circuit (which controls the amplitude of the voltage at the keyer input) is adjustable by the threshold control, R23.

A positive voltage is applied to one plate (pin 2) and to one cathode (pin 1) of V4B and A. Because of the conduction of V4B, C6 effectively charges to its plate potential. After the C6 charge is complete, C7 can become charged through R18 and R19 to the same potential. The positive voltage thus obtained from C7 to ground, in conjunction with the cathode bias developed by R24, determines the operating point of the parallel connected V5 triodes. The value of bias on V5 determines the value of the positive voltage throughout the axis restorer circuit by controlling the plate current of V5A and B, and thereby establishing the drop across the plate resistor, R22.

When an evenly weighted symmetrical signal is coupled through C5 from the discriminator rectifier circuit, V4B rectifies the positive pulses for one instant, and V4A rectifies negative input pulses during the next instant. The charges on C6 and C7 are thus equally displaced in opposite directions.

Because of the difference in resistance values of R18 and R19 (R19 larger), the conduction of V4B during the positive instant caus s a larger value of positive voltage to appear on the V5 grids. The resulting increase in the drop across R22 causes less positive voltage to appear on the keyer (V11) control grid, thus reducing the keyer lock-up voltage.

The threshold control, R23, should be set so that a low level symmetrical signal will produce a bias, or operating axis at V11, which gives the optimum keying. It is the purpose of the axis restorer to maintain this optimum axis for nonsymmetrical signals determined by the conduction of V5, as discussed.

At the grid of V11, a positive voltage of approximately one volt or more is required to swing the flip-flop keyer (discussed later) to key a mark signal. It will remain on mark until a negative voltage of approximately one volt or greater is applied, whereupon it flips over to key a space signal. It will, in turn, remain on space until flipped back to mark by a positive voltage. The grid of V11 is prevented from being driven excessively positive by grid limiting resistor, R21.

The optimum keying axis will depend primarily upon the signal strength and noise conditions. On strong noise-free signals the axis setting is not critical. However, during the reception of weak signals or weak portions of fading signals, a careful setting of the axis is important.

When a nonsymmetrical signal, which has more and/or higher positive pulses (mark) than negative, passes through the system, the axis of the circuit is shifted accordingly. The excess positive signal is rectified by V4B to charge C5, C6, and C7 more positive. This change is reflected through the resistive network comprising R18, R19, R20, and R22, and both sections of V5. As the V5 grid voltage swings more positive, the V5 plate current through R22 increases. The increased drop across R22 subtracts from the increased drop across R20 to restore the axis potential at the V11 grid to that occurring with a symmetrical signal. Note that the increased drop across R20 is the result of the charging action of C5 (up through R20) when a nonsymmetrical signal having a higher positive pulse than negative is coupled through C5.

A signal that is largely space characters (which would ordinarly shift the axis in the negative direction) applies less positive voltage at the V5 grids. The rise in the V5 plate potential again counteracts the shift in the axis.

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Since only the a-c component of the signal reaches the axis restorer circuit from the discriminator, a prolonged mark signal, a prolonged space signal, and a no-signal condition appear the same to the axis restorer, except for the abrupt change at the beginning of the mark or space signal. A mark signal keys a mark output such that a prolonged mark signal will leave the circuit in the mark or no-signal lock-up condition. A prolonged space signal will completely discharge C7, at which time the remaining positive potential across C6 will cause the circuit to switch over to the no-signal lock-up condition.

The time of the lock-up action varies with the setting of the threshold control and with operating conditions. It will range from instantaneous lock-up to a possible condition where two or three seconds are required to lock-up from a strong continuous space signal. The time constant for any one condition is determined by all the components of the axis restorer circuits.

Audio Oscillator-Amplifier

A resistance-coupled audio amplifier comprising tubes V6 and V7 is made to oscillate by the use of positive feedback between the output of V6 and the input of V7. The circuit is that a Wien bridge frequency-selective, resistance-capacity oscillator, which allows one frequency to be applied to the grid of V6 in the proper phase and amplitude to sustain oscillations. The frequency of the oscillations is determined by the values of the components connected in the Wien bridge circuit by S4.

The parallel combination of R26 and the capacitor selected by S4B (C17 through C24 inclusive) form a parallel leg of the bridge network. R25 and the series capacitor selected by S4A (C8 through C16 inclusive) form a series leg of the network.

The regenerative feedback signal is amplified by V7 and applied through C26 to the control grid of the oscillator, V6A. V6B is connected as a diode, which rectifies signals greater than the bias level established at its cathode. The positive voltage is obtained from the R37-R38 junction of the voltage divider connected from B+ to ground. The rectified signal is applied to the V6A grid as bias. This action stabilizes the output level of the oscillator.

The net amplification of the two oscillator tubes is kept low by introducing a fairly large degree of negative feedback together with the positive feedback used to provide oscillations. Assume that the plate voltage of V6A rises slightly. C25 charges to the new potential through R27 and R29 (the cathode resistor of V7.). The increased positive potential at the V7 cathode reduces the feedback amplification provided by the tube, thereby controlling the amplitude of the oscillations. In this manner, harmonics of the oscillator frequency are reduced.

Tuning of the oscillator is accomplished by the use of the frequency-cps switch, S4 in positions one through eight, respectively. When in position nine, the frequency-cps switch allows an external tone from J7 via Z9 to drive the grid of V7. In this position, the oscillator-amplifier functions as a two-stage, resistance-coupled amplifier.

V10 regulates the voltage applied to the plate of the first oscillator tube, V7. The V-R tube holds the voltage to within two volts of the normal 150 volts applied to the plate. The output of the oscillator-amplifier is coupled by C28 to the grid of the phase splitter tube.

Phase Splitter

V8 functions as a conventional phase splitter (paraphase amplifier) circuit to supply push-pull inputs to the grids of the tone modulator, V9. The plate and cathode resistances are approximately equal to insure equal amplitude outputs from these tube elements.

The output from the cathode of V8 is coupled by C30 to the tone modulator, and the plate output is coupled by C29. C31 reduces high-frequency components of the phase splitter output.

Tone Modulator

The tone modulator, V9, is a conventional, push-pull audio output amplifier. The output of the tone modulator is keyed from "tone-on" to "tone-off" by a high, negative bias voltage applied to the grids of the stage from the keyer circuit (discussed later). When the negative signal is applied, the grids of the tone modulator are driven to approximately two times cut-off, which results in no-tone output. When the negative signal from the keyer is not applied, the output of the tone modulator is transform coupled by T3, through the filter, Z7, to the tone output jack, J3.

R50 determines the level (amplitude) of the input signal to the two grids. R51, connected in

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the cathode circuit of both halves of the tube, V9, adjusts the level of the output from the two amplifiers, and is therefore called the output level control.

Keyer

The keyer is a flip-flop circuit employing two triodes V11 and V12, and a neon tube, V13. The neon tube is used for triggering the flip-flop action.

In the mark condition, a positive potential of about one volt (or more) from the discriminator, or axis restorer lockup circuit, is applied to the V11 grid. This causes V11 to conduct heavily, and a large voltage drop is developed across R42 and R46 (in the cathode circuit of V12). The bias voltage of V12 is increased by this action to a point beyond cutoff of the tube. The tendency of V12 cutoff is regenerated at the V12 grid as a result of the plate voltage drop of V11. The action of V11 and V12 is similar to a flip-flop, one-shot multivibrator.

At the cutoff of V12, the plate potential of the tube rises, applying this more positive voltage to the neon tube, V13. V13 ionizes, passing a current through R45 and R59. The voltage drop thus developed across R59 is approximately +37 volts, which is of opposite polarity to that normally applied to R59 from the negative fixed bias supply of -38 volts. In effect then, the voltage output of the keyer circuit is approximately -1 volt, which is applied to the grids of the tone modulator, V9, through R48 and R49, respectively.

Through the action just discussed, the keyer circuit controls the tone-on tone-off condition of the tone modulator. The same action permits the electronic relay (discussed later) to conduct the d-c loop current, which energizes the teletype printer.

A normal space signal (about one volt or more negative) applied to the grid of V11 causes the tube to draw less current. The drop across R42 becomes smaller, allowing the grid of V12 to become less negative. V12 now starts to conduct, and because of the common cathode coupling between V11 and V12, the cathode of V11 becomes more positive, causing this tube to stop conducting.

When V12 starts to conduct, a larger voltage drop is developed across R45, which reduces the voltage at the trigger tube V13. The ionization of the trigger tube ceases, and the -38 volts normally applied from the power supply again takes control of the tone modulator and electronic relay circuits. Under this condition, both the tone modulator and electronic relay circuits block the passage of plate current in their respective stages. This results in no tone from the tone modulator, and the electronic relay opens the d-c loop circuit that energizes the teletype printer.

When the input to V11 swings from approximately 1 volt positive to 1 volt negative, or wider, the keyer produces sharply defined characters at the output of the unit. This occurs as a result of the regenerative action in the circuit comprising R46, V11, and R43, as discussed earlier.

Electronic Relay

The electronic relay employs tubes V14 and V15, which act to open and close the d-c loop of the teletype printer. The tubes are connected in parallel and obtain their plate supply voltage (approximately 120 volts) from the teletype printer loop supply. For this reason, it is necessary to directly ground the cathodes of V14 and V15, because the negative side of the d-c and loop circuit is already at ground potential. R57, R60 are parasitic suppressors to counteract the tendency of parasitic oscillations.

The electronic relay is keyed opened or closed by the keying voltage from the flip-flop keyer circuit, as discussed. The voltage is applied at the grids of V14 and V15 along dual paths from the R59-V13 junction.

Under the mark condition, the keying voltage is about one volt negative, which allows the electronic relay tubes to conduct, closing the d-c loop of the teletype printer. In the space condition, the keying voltage is approximately 38 volts negative. This completely blocks the conduction of V14 and V15, opening the circuit of the d-c teletype loop.

COMPARATOR

Mark-Space Selector Circuit

Two double diodes, V16 and V17 (fig. 11-5) are connected together in a special circuit for automatically comparing the two diversity signals. The circuit selects the better mark and space pulse for each character of the coded message. A four-position selector switch, S6, chooses the mode of operation of the comparator. The positions correspond to (1) tune, (2) channel A, (3) combined, and (4) channel B.

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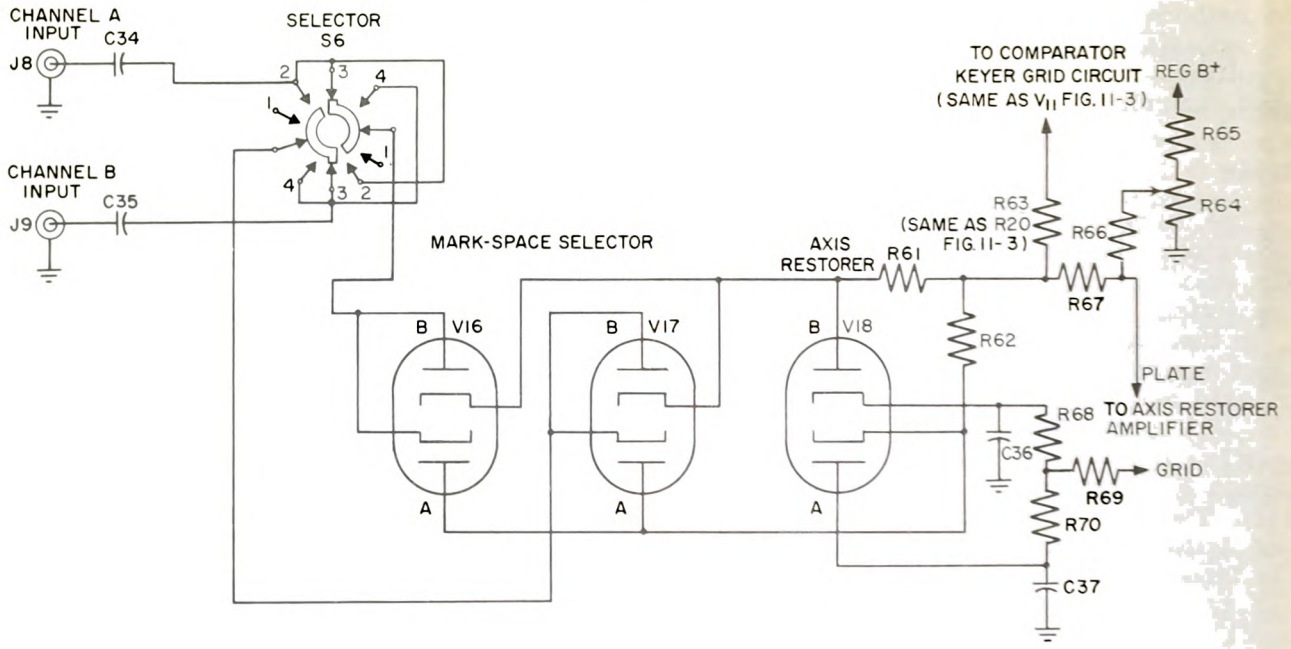

Figure 11-5. —Comparator mark-space selector circuit.

Most of the circuits of the comparator are identical to those of the converter, just discussed. Only those circuits appearing in figure 11-5 differ sufficiently to warrant separate discussion.

When S6 is placed in the DIVERSITY position, 3, (as shown), the signal from channel A converter is applied through C34 and S6 to the plate of V16B and the cathode of V16A. During positive input pulses, if V16B conducts, the circuit will be through S6, C34, and J8, through the shorted contacts of S3 (fig. 11-3) and the Z5 filter to ground. From ground, the circuit is completed through the threshold control, R64, of the comparator, R66, R67, and through R61 to the cathode of V16B. Negative input pulses cause a similar conduction of V16A through R62, R63, R66, and R64 to ground, and through the external circuit to return through C34 to the V16A cathode.

The signal from the channel B converter is similarly connected to V17B and V17A to pass the positive pulses through R61, and the negative pulses through R62.

To understand the comparing and selecting action, consider that the first converter was delivering a three-volt positive pulse at the instant the second converter delivers a two-volt similar pulse. The three-volt pulse of V16B passes through the diode with negligible drop to produce a three-volt potential across its cathode resistors. This will result in a three-volt drop at the cathode of V17B, and there is only a two-volt pulse at its plate. Thus, only the diode with the higher potential in the conducting direction will pass the signal. This same selection occurs for negative pulses at V16A and V17A, with R62 acting as the common load.

When the two signals have pulses of equal value, there is some combining (noise particularly) because of phase differences, and the resultant signal is passed. Otherwise, the circuit passes only the stronger positive, or mark, pulse, and the stronger negative, or space, pulse. The selection is instantaneous, even to the point of selecting a portion of a poorly shaped pulse.

The output of the mark-space selector circuit is taken from the R61-R67 junction and applied through R63 to the keyer circuit. The d-c potential at the output junction is determined by the axis restorer amplifier circuit and the threshold control setting. The signals developed by R61 and R62 do not cancel since they do not occur simultaneously.

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Axis Restorer

The axis restorer circuit in the comparator is identical in circuitry and function to that discussed in the converter, except for the circuit into the axis restorer diode, V18A and V18B. V18B receives the positive mark voltage at its plate and rectifies this portion of the signal input. Likewise, the circuit of V18A receives and rectifies the negative portion of the input. The two diode sections pass the signal to C36 and C37, respectively. The resulting action in biasing the axis restorer circuit through the use of the signal pulses is the same as that discussed in the converter axis restorer.

The d-c component in the output of the mark-space selector circuit is not isolated from the following signal circuits by a capacitor, as is done from the discriminator in the converter. The axis restorer is definite enough in its control not to be appreciably affected by this voltage. Moreover, the voltage thus derived establishes the bias at the R61-R62 junction into which the mark-space selector diodes operate.

Power Supply

The power supply (not shown) is a conventional full-wave rectifier, which provides the power required by the comparator circuits. It furnishes +250 volts d-c at 35 ma, and has a separate rectifier circuit that supplies the required 32 volts negative bias.

TONE TERMINAL AN/SGC-1

The AN/SGC-1 (fig. 11-6) is a teletypewriter terminal for the transmission and reception of teletypewriter messages between communication stations. During transmission the terminal set converts the current pulses from the teletypewriter equipment into audio tones that modulate the radio transmitter frequency. When receiving, the equipment converts the audio tones to make-and-break signals for operating the associated teletypewriter.

Conversion of the current pulses to audio tones is accomplished by an audio oscillator in the tone converter that operates at 700 cps for a marking signal (closed d-c loop) and at 500 cps for spacing signals (open d-c loop). An internal relay of the converter closes a control line to the radio transmitter, which places the unit on the air at the instant the operator begins typing the message. The control line remains closed until after the message has been transmitted.

When receiving messages, the converter accepts the mark and space tones incoming from the associated receiver, and passes the intelligence of the tones to the make-and-break contacts of a relay connected in the local teletypewriter d-c loop. The teletypewriter prints the message in accordance with the make-and-break impulses.

THE RECEIVER CIRCUIT BLOCK DIAGRAM

An input filter of the tone converter (fig. 11-7) provides an attenuation of frequencies from 400 to 800 cps not greater than 2 db. The filter functions to attenuate other frequencies as much as 40 db. An attenuator switch permits adjustment of the degree of attenuation. In addition to the attenuating characteristics, the filter provides a circuit that aids in matching the converter terminal to the 600-ohm impedance input.

The amplifier-limiter circuit provides a constant signal level to the remaining circuits of the converter. Its output is coupled to the frequency discriminator filter, which provides separate circuits for the selection of the mark (700 cps) and space (500 cps) frequencies.

Two germanium rectifiers at the output of the frequency discriminator filter rectify and convert the respective frequencies to a corresponding d-c voltage. This voltage then serves as the input to the individual direct current amplifiers. The amplifiers are connected so as to cause the receive relay to close its contacts when the incoming signal is 700 cps, and to open its contacts when the incoming signal is 500 cps.

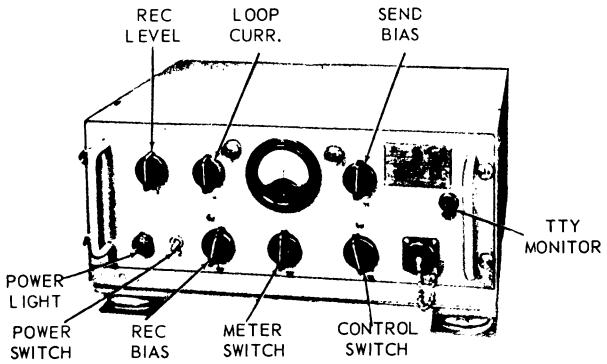

Figure 11-6. —Radio teletype tone terminal AN/SGC-1.

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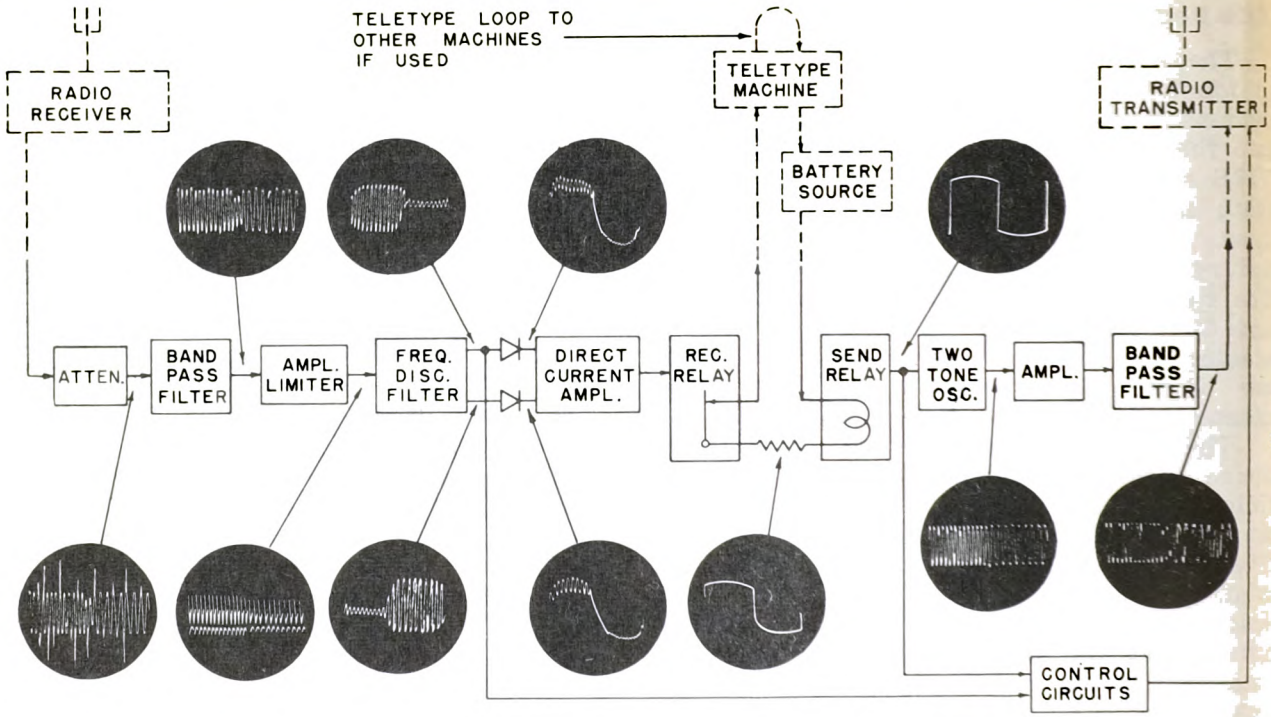

Figure 11-7. –Tone terminal, block diagram.

One set of the contacts of the receive relay is placed in series with the local teletypewriter loop. Consequently, the receive relay is able to open and close the local teletypewriter loop, thus forming current pulses in the loop identical to those that originated at the transmitter.

Since the receive and send circuits of the tone converter will be discussed separately, the remaining portion of the block diagram (fig. 11-7) will be discussed later in this section. Reference to this diagram will be made at the appropriate point in this discussion.

OPERATION OF THE RECEIVE CIRCUITS

The audio signals (fig. 11-8) of 500 and 700 cps, respectively, are coupled from the receiver (preferably through a 600-ohm line) to the tone converter input by T1. E1 represents a variable attenuator-bridge, wire-wound T section, which allows a 20-step attenuation of 2 db per step, thus giving a possible maximum attenuation of 40 db for the stronger signals. The degree of attenuation is determined by the setting of the receive level control on the converter front panel (fig. 11-6).

A six-db pad comprising R1, R2, and R3 provides additional attenuation, and minimizes the effects of mismatch if the output impedance of the associated receiver is not 600 ohms. The signal then passes through the control switch, S3, in the AUTO or REC STBY positions (contacts 8 and 9) to the input filter, Z1.

Frequencies from 400 to 800 cps are attenuated no more than 2 db by the input filter. Frequencies above or below this range are attenuated as much as 40 db. The Z1 output is applied to the T2 primary.

Amplifier-Limiter

The signal from the T2 secondary is applied through grid limiting resistor, R4, to the first amplifier-limiter (V1A) control grid. Upon reception of weak signals, the stage functions as a conventional amplifier, applying its output through C3 and R10 to the second amplifier-limiter, V1B. The signal is subsequently coupled (in the same manner) to the V2A and V2B control grids. The operation of all the stages is identical, except V2B, which will be discussed later.

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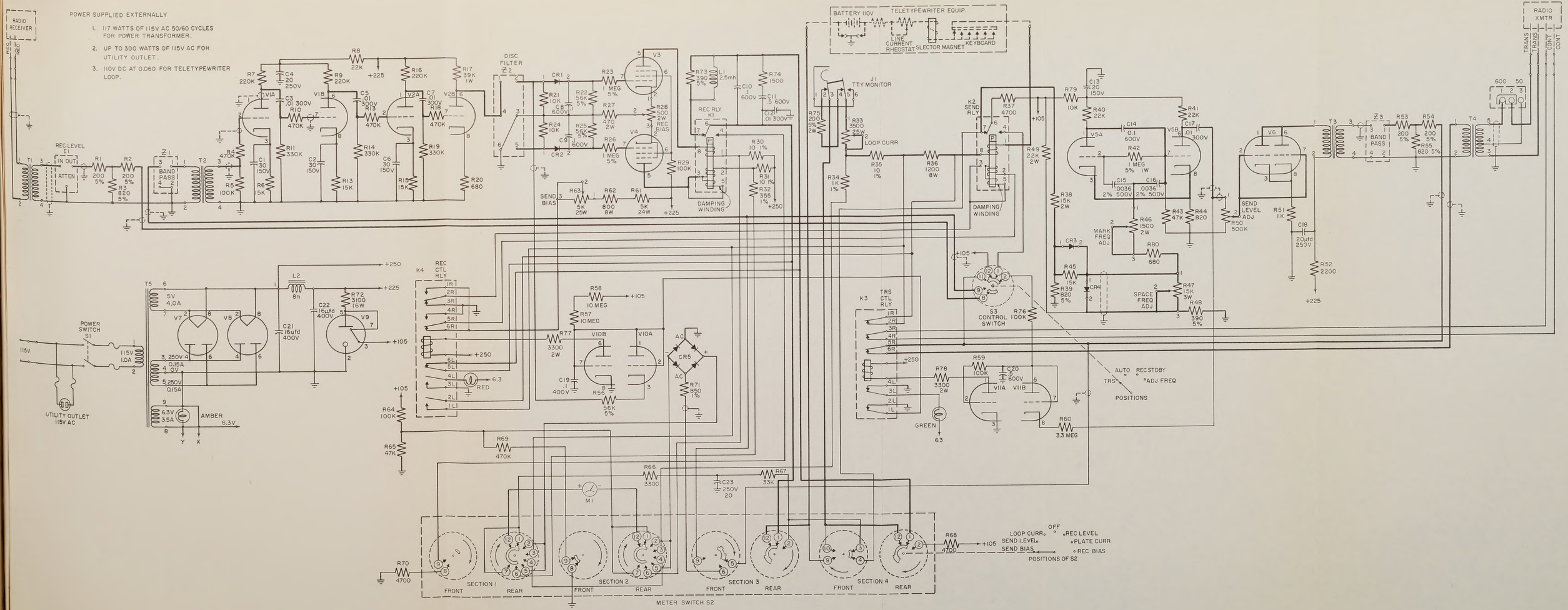

Figure 11-8. –Tone terminal, schematic diagram.

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The plate load resistors of the respective stages (R7, R9, R16, and R17) are made large to facilitate grid limiting. On strong positive peaks, the grids of the tubes draw a current through their grid resistors, R4, R10, R13, and R18, respectively. The voltage drop developed across these resistors is such that the potential at the grid (negative) will oppose the positive going input, and grid limiting occurs on this portion of the positive alternation. Strong negative input signals drive the grids to cutoff, which limits the negative alternations. The cathode resistors are made large so that they operate the tubes in the mid-portion of their characteristic curves.

The fourth stage, V2B, utilizes different circuit constants in its plate and cathode to provide proper impedance matching for the discriminator filter circuit that follows. To correct for possible discrepancies in the positive and negative halves of the amplified (or limited) signal, degenerative action is accomplished across R20 in the V2B cathode.

Frequency Discriminator Filter Circuit

The frequency discriminator filter circuit, Z2, consists of two complimentary bandpass filters in one container. Terminals 1, 2, and 3 connect to the section that passes 700 cps and a narrow band of frequencies above and below that center frequency. Terminals 3, 4, and 5 connect to the section that passes a similar band centering on 500 cps. Terminals 3 and 6 are connected together and grounded.

The input terminals of the two sections (terminals 1 and 4 respectively) are tied together and connected to the plate of the last amplifier-limiter, V2B. The internal capacitors of the filters (not shown) eliminate the need of a coupling capacitor.

The two incoming signals from the receiver may be at different amplitude levels. However, the action of the amplifier-limiter causes the signals to arrive at the discriminator filter circuit at a constant level. The output of the filter is either 500 or 700 cps with respect to the time of the mark and space signal input. The two frequencies do not appear at the filter output simultaneously. A small amount of leakage exists between the two filters, but not in sufficient amount to cause any appreciable harmful effect.

The mark tone, 700 cps, appears across terminating resistor R21, and the space tone, 500 cps, appears across R24. During the time the mark tone is being applied across R21, CR1 conducts to charge C8 (on positive half cycles) to approximately 15 volts. During negative half cycles of this input, C8 loses some of its charge through R22. As the discharge resistor, R22 is about six times as large as the charging resistance of CR1, the majority of the accumulated charge is retained by C8 until the end of the mark tone. At this time C8 will discharge to the zero potential.

During the time the space tone is being received, the signal charges C9 through CR2 and R24. C9 likewise discharges on negative half cycles through R25. As a result of the discharge of C8 and C9 during mark and space intervals, the d-c voltage developed across R22 and R25, respectively, is applied to the individual control grids of V3 and V4.

D-C Amplifier

The cathodes of the d-c amplifier pentodes, V3 and V4, are connected to opposite ends of the receive bias control, R28. The arm of the control is connected to the common cathode resistor, R27, so that it, too, is common to both tubes. R28 tends to equalize the two cathode currents, as discussed later. The screen voltage of both tubes is maintained at the same value by the action of R29, which tends to provide equal plate currents from V3 and V4.

Mark impulses applied through R23 to the grid of V3 cause the plate current to increase from approximately 2 ma to 22 ma. Space impulses applied through R26 to the grid of V4 cause a similar increase in the plate current of V4. The receive bias potentiometer, R28, permits changing the relative amplitude of these impulses by varying the magnitude of the pentode plate currents.

When the receive bias control is rotated in one direction, it increases the resistance in one cathode circuit while decreasing the resistance in the other. This action causes the cathode voltage of the first pentode to increase and causes the cathode voltage of the second pentode to decrease. Since the plate current of the pentodes is controlled by the grid-to-cathode voltage, a change in the cathode voltage by varying the receive bias potentiometer is very effective in changing the plate current of the two pentodes.

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Receive Relay Circuit

Relay K1 is a polar relay containing a permanent magnet and three controlling coils. The plate current of V3 and V4, respectively, is caused to pass through separate windings of the relay. The moveable arm of K1 (contact 6) is operated under the influence of the coils to contact 7 for a mark input from the d-c amplifier, and to contact 4 for the spacing signal.

Current from V3 passes through the K1 coil (pin 8 and 1) and meter shunt R30 to the B supply. The current through the 8 and 1 coil of K1 causes the arm of K1 (contact 6) to make with contact 7 during mark. This action completes the teletypewriter loop from contact 7 of K1, through the keying filter comprising R73 and L1, through contacts 1 and 12 of S2 (section 4 rear), through J1 contacts 3 and 2 and R75 (compensating resistor), through contacts 4 and 5 of J1 and the loop current adjust, R33, through meter shunt R35 and 2L and 1L of K4, through coil 8 and 1 of the send relay, K2, through the local teletypewriter keyboard and selector magnet, through the line current rheostat and power supply, through contacts 12 and 1 of S2 rear (section 3) to return to contact 6 of relay, K1, thus completing the loop. Note again that the received signal passes through both the receive and send relays.

The spacing current from V4 passes through coil 3 and 2 of K1 to move contact 6 of the relay to make with contact 4. In this condition, the teletypewriter loop is open.

L1, R73, C10, C11, R74, and C12 collectively form a spark suppression and waveshaping network. The network compensates for the inductive reaction peculiarity associated with some models of teletypewriters.

The third winding consists of only a few turns, which are shorted at the socket connections. During the time that the 8-1 and 3-2 coils produce collapsing fields, the shorted coil builds a field, which counteracts the changing flux. This action damps the coils to prevent possible oscillations. The damping permits high-speed keying without contact bounce.

R30 and R31 are used for measuring the plate current of V3 and V4, respectively. The voltage developed across these resistors is proportional to the plate currents, and is applied to the meter through R32 and S2 (contacts 6 and 12 of section 1 rear) when the meter switch is in the PLATE CURRENT position.

Operation of the Receive Bias Adjustment

All impulses from the teletypewriter keyboard, with the exception of the stopping impulse, should have the same duration. In the 7.42 code system, the duration of the impulses corresponds to 22 ms for the starting space, 22 ms for each portion of the 5-coded characters, and 31 ms (42% longer) for the stopping mark.

In the process of converting the intelligence from d-c pulses to tone, or any of the various methods used for communications, a distortion known as bias is often introduced. This distortion results in the mark and space impulses being elongated or shortened with respect to their normal length. The length of time for a teletypewriter letter (discussed later) is fixed by the mechanical gearing of the machine and cannot be altered or distorted. However, the individual mark impulses can be lengthened, while the space impulses are shortened simultaneously, or vice versa. If the marking impulses are longer than normal, the condition is known as marking bias. If the spacing impulses are too long, the condition is known as spacing bias.

A teletypewriter can operate with a considerable amount of bias in its signal. Exactly how much bias it can stand without misprinting is determined by such things as the adjustment of its mechanism or the speed of its motor. These items are discussed later in this chapter.

To compensate for bias, the receive bias control must be properly adjusted. As considered earlier, the operation of this control caused the plate current of V3 and V4 to vary accordingly. Figure 11-9, A shows the shape of the mark and space impulses when the incoming tones are evenly spaced groups of 700-cps and 500-cps signals. For the purpose of explanation, the space impulses are shown upside down to indicate the opposing action of the two plate currents upon the resultant flux in the receive relay. The ripples in the waveforms are due to the discharge of C8 and C9, as discussed. They present no appreciable effect on the operation of the relay.

The ripples show that the space tone starts as soon as the mark tone ends, and vice versa. The space-to-mark transition (S-M) closely follows the starting time of the mark tone. Likewise, the mark-to-space transition (M-S) closely follows the starting time of the space tone. These transitions take place immediately after the two plate currents of V3 and V4 are equal.

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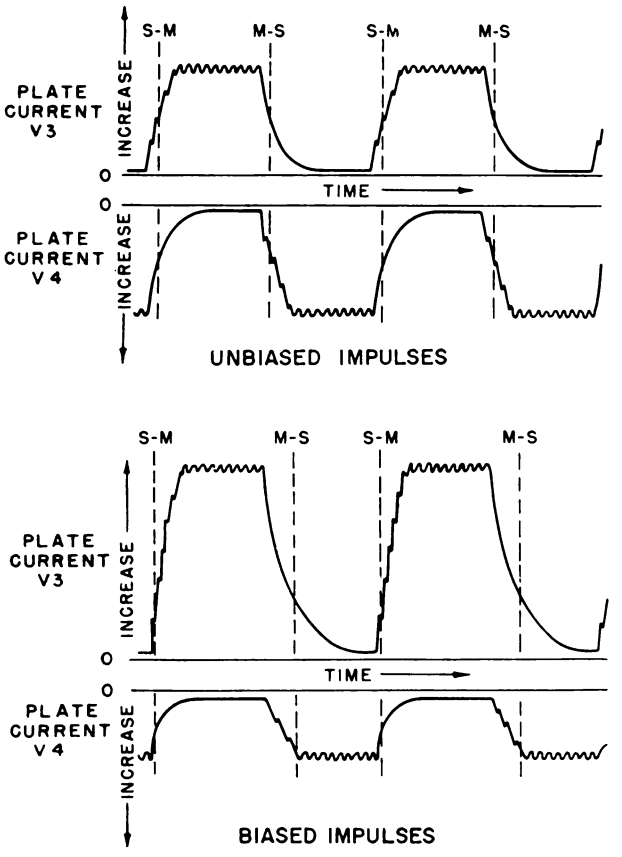

Figure 11-9. —Biased and unbiased teletype impulses.

Figure 11-9, B, shows the effect when one plate current, V3, is increased and the other plate current, V4, is decreased by the operation of the receive bias control. The ripples show that the timing of the incoming tones has not changed. The S-M and the M-S transitions still take place when the two plate currents are equal, but, it should be noted that the relative spacing on the timing axis has changed. As the length of time from the S-M to the M-S transition has increased, the relay armature would be operated to, and held against, the mark contact longer than normal, and the d-c signals to the teletypewriter would have marking bias. Rotation of the receive bias control in the opposite direction would cause the above conditions to change conversely, and the signals to the teletypewriter would have spacing bias.

Under actual conditions, it would be undesirable to introduce either marking or spacing bias, and this explanation is intended only to illustrate how the receive bias control can compensate for a bias distortion that is introduced into the signal either internally or externally. Where the timing of the tone groups is distorted, or the amplitude of the rectified mark and space impulses is incorrect, the receive bias control will usually permit sufficient compensation to correct the signal.

TRANSMIT CIRCUIT BLOCK DIAGRAM

The send relay circuit (fig. 11-7) energizes the transmit circuits when a message is to be transmitted. In the receive condition, the send relay is deactivated.

A single, two-tone oscillator (multivibrator) provides the two tones of 500 and 700 cycles, respectively. A mark and space signal is represented at the oscillator output in the diagram.

Signals from the two-tone oscillator are amplified and applied through a bandpass filter to the associated transmitter. The filter passes frequencies from 400 to 800 cps with little attenuation. However, frequencies outside the filter range receive a correspondingly greater amount of attenuation with an increase in frequency on either side of the bandpass of the filter.

Signals from the frequency discriminator filter and the two-tone oscillator are applied to the control circuits, which include a transmit and receive control relay and their associated amplifier tubes (discussed later). These control circuits exert a direct influence on the transmit and receive circuits to provide the automatic switching action of the tone terminal.

OPERATION OF THE TRANSMIT CIRCUITS

Send Relay Circuit

The send relay, K2, (fig. 11-8) is also a polar type relay with three coils. The current in the 8-1 winding (connected in the teletypewriter loop) is 60 ma in the mark condition and 0 ma in the space condition. In standby condition (neither receiving nor transmitting) the current in the 8-1 winding is reduced to 36 ma. Though this action is described in detail later, note that R36 (in series with the teletypewriter loop) causes the current to rise or fall from 60 to 36 ma, as determined by the shorted or open condition of the 1R-2R contacts of K3.

A current flowing through the 3-2 winding and the voltage dividers R63, R62, and R61, is

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steady at 30 ma. The send bias control, R63, provides a method of adjusting this bias between 20 and 40 ma. The second winding (3-2) is hereafter referred to as the space winding.

In the standby condition, the 30-ma current in the space winding (3-2) produces a magnetic flux that is weaker and made opposite to that flux produced by the 36 ma closed teletypewriter loop current in the 8-1 winding. The resultant flux holds the send relay armature (6) against the mark contact (7). The third winding produces a damping action in the same manner as presented in the discussion of the receive relay, K1.

Although the above discussion is relative to the standby condition, the same action of the coils is repeated during the marking condition, except that the loop current is increased to 60 ma. For space, the teletypewriter loop current drops to 0, and the unopposed flux produced by the space winding (30 ma) moves the relay armature (6) to the space contact (4). When a message is typed, the loop current is rapidly and repeatedly changed from 0 to 60 ma and back to 0 again. The resultant flux is continually reversed, and the relay armature is continually switched from mark (pin 7) to space (pin 4) and back again.

In the receive condition the 30-ma current in the space winding is reversed by the action of the receive control relay, K4, which operates to make 4R-5R and 2R-1R contacts. Thus, the flux of the space winding is reversed such that it now aids the flux produced by the teletypewriter loop current. Therefore, regardless of the loop current, the resultant flux of the send relay is always in the same direction. The reversal of the space winding current constitutes the interlocking feature that prevents the send relay from operating when receiving a message.

Receive Control Relay

The twin triode, V10, has one of its sections, V10A, connected as a diode. One end of the receive control relay is tied to the 250-volt B supply, and the other end is connected through current limiting resistor R77 to the plate (pin 6) of the V10B triode. The relay is energized when plate current flows. The plate current is controlled by the bias voltage on grid 7 of V10. In general, two factors determine the bias applied at this grid. The grounded or ungrounded condition of the transmit control relay (contact 1L and 2L) is the first determining factor, and the presence or absence of an incoming signal at the V10A cathode is the second.

In standby, no signal is received, and the transmit control relay (contacts 1L and 2L) connected to the grid of the triode, is open. Under these conditions, C19 assumes a positive charge from the +105-volt supply. In the voltage divider comprising R58, R57, and the grid-to-cathode resistance of V10B, R58, and R57 assume more than 99% of the voltage drop. This limits the cathode-to-grid positive voltage to less than one volt. With this grid potential, V10B conducts heavily through R77 and the solenoid of K4, to operate its contacts.

Suppose now, that the receive circuit begins to receive a signal. The mark frequency taken from the frequency discriminator filter output is applied through R56 to the V10A cathode. Negative half cycles of the input cause V10A to conduct. The circuit containing R56, V10A, and C19 form a much shorter time constant (when V10A conducts) to charge C19 negative than does the circuit comprising C19, R57, and R58, which tries to charge C19 positive. As a result, the higher current in the lower time constant circuit (when V10A conducts) charges C19 negative with respect to ground. The charge is cumulative on alternate half cycles of the signal until C19 assumes a sufficient charge to cut off V10B. This action deenergizes K4 to place it in the receive condition. A short time after the last mark cycle (ending the receive condition) C19 completely discharges and assumes a positive charge. V10B again conducts to energize the receive control relay, K4. The time of the discharge of C19 is determined by the time constant of C19, R57, and R58. This circuit forms a long time constant discharge path.

In the transmit condition the transmit control relay, K3, is energized, as discussed later. Energized K3 shorts the grid-to-ground potential developed by C19. The cathode and grid of V10B are therefore at ground potential, allowing plate current to flow through the K4 solenoid. Thus, the receive control relay is energized during transmission. Since the transmit control relay, K3, keeps both ends of C19 at ground potential, any incoming signal from the frequency discriminator filter cannot charge C19 in an attempt to deenergize the receive control relay.

Only in the receive condition will contacts 1R-2R and 4R-5R of K4 be closed. The 1L-2L contacts of the transmit control relay, K3, are open at this time. This action causes the send

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relay space winding current to be reversed. Thus, contacts 7 and 6 of the send relay, K2, are constantly held closed throughout the receive condition, as discussed. Note that the receive control relay, K4, is always energized except when the terminal is in the receive condition.

Contacts 5L and 6L of K4 are closed during transmit and standby (as shown), shorting pins 6 and 7 of the receive relay, K1, so that the teletypewriter loop is closed at this point. Contacts 3L and 4L supply 6.3 volts to the receive indicator lamp when in the receive condition. Contacts 1L and 2L short R36 (in the teletypewriter loop) in the receive condition, allowing the loop current to rise from 36 to 60 ma through the K1 solenoid during mark.

Transmit Control Relay

V11 functions in much the same manner as V10 to control the action of the transmit control relay, K3. Plate current passes through the K3 solenoid when V11A is conducting. The grid bias of V11B is determined by the position of the control switch, S3, and the position of the send relay, K2.

The cathode of the diode section of V11B is connected through R60 to the output of the audio oscillator, V5 (discussed later). The oscillator functions continuously upon the application of the B supply voltage, and a continuous audio voltage is constantly applied to the V11B diode.

In the transmit position of the control switch, S3, (as shown) mark and space signals in the teletypewriter loop pass through the solenoid of the send relay, K2. A +105 volts from the voltage regulator, V9, charges C20 positive with respect to ground through R76 and the S3 contacts. The positive potential of C20 is applied to the control grid (pin 2) of V11A, allowing the tube to pass an energizing current through the solenoid of K3.

The 1R-2R contacts of K3 short R36 in the teletypewriter loop to allow the 60-ma mark current. Contacts 3R and 4R close the control line to the transmitter during transmit condition, which places the transmitter on the air. Contacts 5R and 6R close the output circuit during transmit condition so that the outgoing signal tones may be applied across the T4 primary.

The 3L-4L contacts complete the circuit to the transmit indicator lamp, and contacts 1L-2L ground the grid of V10B, which causes it to operate the receive control relay as discussed.

With the control switch, S3, in the AUTO position, an incoming signal causes the terminal to change from standby condition to receive. Keying the local teletypewriter causes the terminal to change from standby condition to transmit.

In standby and receive condition, the send relay, K2, connects its armature (6) to the mark contact (7) so that the space contact (4) is open. In this condition, the only voltage applied to the V11B diode is the oscillator a-c voltage. On negative half cycles of this input, V11B conducts to charge C20 negative with respect to ground. This negative voltage is applied through grid limiting resistor, R59, to the V11A grid. V11A cuts off, thus preventing the flow of plate current through the solenoid of K3. This relay is therefore deenergized during standby and receive conditions.

Since K3 is not energized, its contacts 1L and 2L are open. The incoming signal at the V10A cathode cuts off V10B and deenergizes the receive control relay, K4. Thus, contacts 4R and 5R complete the space winding circuit of K2 such that its current is reversed, and the armature of the send relay is held in the mark or receive condition.

When a message is being typed locally (S3 in the AUTO position), the armature of the send relay intermittently applies positive 105 volts to the space contact (4). This voltage charges C20 through R76 and contacts 1 and 2 of S3. The positive voltage thus developed across C20 is applied through R59 to the V11A triode, and K3 is energized. Contacts 1L and 2L of K3 are now closed, and V10B conducts to energize K4. The current through the send relay now changes back to its original direction since the 1R-2R contacts of K4 are again closed. This allows the armature to move between pins 7 and 4 under the influence of the teletypewriter characters.

In the ADJ FREQ position of the control switch, S3, all contacts are open except contacts 9 and 11. These contacts connect the output circuit of the terminal to the input so that the mark and space tones produced by the oscillator are fed into the receive circuit. The receive circuit separates the mark and space tones and applies rectified signals to the d-c amplifier in the same manner as discussed. When each tone is at its correct frequency, it receives a minimum of attenuation in passing through the discriminator input filter. Therefore, when the signal is rectified, it produces a maximum

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voltage at its corresponding frequency. With maximum voltage applied to the d-c amplifier, the circuit produces a maximum of plate current. With the circuit connected in this manner, the mark and space frequency adjust controls (discussed later) can be set for optimum operation.

Two-tone Oscillator

The circuit of V5 is designed to produce the 700- and 500-cps mark and space frequencies used to modulate the frequency of the associated r-f transmitter. Both the mark and space frequency adjust controls are provided to exact the respective frequencies.

Disregarding the circuit comprising R42, C15, and C16, and completing the V5A grid circuit, the oscillator can be described as that of a conventional cathode-coupled multivibrator. If we, therefore, assume V5A to be conducting, C14 discharges through R43; the common cathode resistor, R44; and the conducting resistance of V5A, to return to C14. The negative voltage across R43 toward the V5B grid cuts off this section of the tube until after the complete discharge of C14. At this time, with the plate voltage of V5B at the B supply value and zero voltage across R43, V5B begins slight conduction.

Since the conduction of V5A has developed a large fixed bias voltage across R44, the current through R41 is small at the beginning of the V5B conduction. However, the small conduction of V5B is sufficient to raise the R44 bias voltage, which, in turn, moves the operating point of V5A down on its characteristic curve. The rise in plate voltage of V5A charges C14 through R43, thus aiding the V5B conduction. The latter tube section then moves almost instantaneously up its characteristic curve to the saturation point. The circuit will remain in this condition until a positive trigger pulse is applied to the V5A grid.

The null network comprising R42, C15, and C16 couples all but one of the harmonic frequencies from the V5A plate through C14 to the V5A grid. The circuit in effect cancels one frequency or presents a null to the V5A grid at this frequency.

All frequencies passed by the network appear at the grid, and the same signals simultaneously appear at the cathode as a result of plate current flow. Signals applied to the grid and cathode at the same time and 180° out of phase will not affect the plate current of the tube. These frequencies are therefore degenerated. Since the null frequency is only applied at the cathode (caused by plate current through R44) it has effect on the plate current. The null signal voltage across R44 acts as the trigger input for V5A. Grid leak components C14 and R43 have very little effect on the frequency of the oscillator.

The null frequency (f_n) is a function of the size of components R42, C15, C16, R46, R80, R47, and R48. The latter two resistors are paralleled by R39 and R45. The formula for determining the oscillator frequency is

$$F_n = \frac{1}{2\pi\sqrt{R42\ Re\ C15\ C16}}, \text{ where}$$

Re is the sum of R46 and R80, in series with the parallel combination of R39-R45 and R47-R48 to ground. With all the resistors in the circuit, the oscillator functions at the lowest frequency.

Send relay K2 changes the oscillator frequency. To increase the frequency to 700 cps, the resistance must be decreased as indicated by the formula. This is accomplished by shorting R47 in the following manner. A potential of approximately 2 volts positive is developed across R48 as a result of a current flow through R48 and R49 to the +105-volt supply. This establishes a +2 volt-potential at the bottom end of CR4. When K2 is in the mark condition (contact 7), another current flows from ground through R39, R38, the send relay contacts (7 and 6) and R37, to the +105-volt supply. Approximately +4 volts is developed across R39, which is applied to the arrow of CR3. The two diodes, CR3 and CR4, therefore conduct during mark condition to short R47. Thus, the null network resistance is decreased, and the oscillator frequency rises to 700 cps. With R46, R80, and R48 in the circuit, the mark frequency can be adjusted to excatly 700 cps by the mark frequency adjust control, R46.

During space condition, armature 6 of K2 is disconnected from the +105 volts at its contact 7. The two diodes (CR3 and CR4) do not conduct as a result of the +2 volts still applied to the bottom end of CR4. In effect then, the diodes are not in the circuit, and the full resistance of R47 is again placed in the null network circuit. Likewise, the oscillator frequency decreases to 500 cps. The proper setting of the space frequency adjust, R47, at this time will correct the space frequency output.

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Tone Amplifier

Mark and space tones from the oscillator at the V5B plate are coupled by C17 to the tone amplifier, V6. This amplifier comprises a single stage twin triode with the two sections of the triode parallel connected to provide low plate resistance. The input level of the tones is adjustable by means of the send level potentiometer, R50, in the grid circuit. R52 and C18 decouple the plate of V6 from the B supply. Degeneration, because of the unbypassed cathode resistor, R51, aids in maintaining the two frequencies at approximately the same level. The amplified tones are coupled to the output circuit through the plate transformer, T3.

Output Circuit

The signal at the T3 secondary is applied to a bandpass filter, Z3. The filter passes frequencies from 400 to 800 cps with an attenuation not greater than 2 db. Frequencies outside this range receive greater amounts of attenuation dependent upon its harmonic frequency.

The filter output is applied to a 6-db pad comprising R53, R54, and R55. This circuit is identical to the 6-db input pad and functions to minimize the effects of a possible mismatch between the terminal and radio transmitter.

Before the signal can reach the impedance matching transformer, T4, it must pass through the 5R-6R contacts of the transmit control relay, K3. These contacts are closed when a message is being transmitted. The secondary of T4 is tapped for matching to either a 600- or 50-ohm output impedance.

METERING CIRCUITS

Meter M1 is a d-c (zero center) milliammeter requiring a current of 1 ma for full scale deflection. 100-0-100 and -10 to +5 dbm scales are shown on the meter. The meter is switched into the circuit by the meter switch, S2.

In the SEND-BIAS position of S2 the send bias control, R63, may be adjusted so that the meter reads zero, indicating proper length of the mark and space tones being transmitted. Switch contacts 8 and 9 of S2 (section 2 front) are closed in the SEND BIAS position to parallel R69 and R65. This connection places a positive potential (of a predetermined value) at the negative side of the meter, M1. The positive side of the meter is connected through a current limiting resistor, R66, and a filter section comprising C23 and R67. The filter section is connected to the mark contact (pin 7) on the send relay, K2.

The third impulse of a spacing signal represents a current in the loop, and the first, second, fourth, and fifth impulses are no-current signals. If the space bar on the teletypewriter keyboard is held down, the send relay armature moves back and forth between its mark and space contacts. A positive 105 volts is applied to the mark contact (7) from the B supply when the armature is in the mark condition. This positive voltage is applied to the metering circuit. When the armature is on the space contact (4), the positive voltage is not applied to the metering circuit.

On mark impulses, C23 acquires a charge from ground through R67 and R37. During the spacing impulses, C23 may discharge through R65 and R69 (in parallel) through the meter and R66 to the positive side of C23. Because of the fast movement of the send relay armature, the voltage on C23 stabilizes at a value that is dependent upon the relative length of the mark and space impulses. If the space impulse is too long, the voltage on C23 has more time to leak off, and the voltage applied to the meter from C23 is low. Conversely, if the space impulse is too short, the voltage applied to the meter is too high. The result is that when the mark and space impulses are of proper relative length, the C23 voltage and the positive voltage applied to the negative side of the meter are the same. In this condition, the meter reading will be zero.

In the SEND LEVEL position of S2 (one clockwise rotation of all sections) the send level potentiometer, R50, may be adjusted so that the meter indicates the proper level of the signals being fed to the transmitter. R50 controls the level of the signal applied to the tone amplifier, V6.

With S2 in the SEND LEVEL position, the output of the bandpass filter, Z3 (passing through the output 6-db pad) is applied through section 3 front of S2 (contacts 5 and 6) through R71 to a bridge rectifier, CR5. The rectifier changes the signal to d-c pulses, which are applied through section 2 rear of S2 (contacts 2 and 12) to the meter, M1. The circuit path is completed through contacts 12 and 2 of section 1 rear, the bridge circuit, CR5, and ground, to return to the filter, Z3.

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The LOOP CURRENT position of S2 switches the proper components for metering and adjusting the loop current. With the switch in this position, the R34-R35 junction is connected by S2 to the negative side of M1 through contacts 3 and 12 of section 2 rear. Section 1 rear (contacts 12 and 3) connects the positive side of M1 to the R35-R36 junction. In this manner, the meter and multiplier resistor, R34, are placed across R35.

The voltage developed by R35 is proportional to the current flowing in the teletypewriter loop circuit. Loop current rheostat R33 can now be adjusted so that the meter reads 60 on its upper scale when the terminal is in the transmit condition.

In the OFF position of S2, contacts 4 and 12 of section 2 rear and contacts 4 and 12 of section 1 rear short the meter, M1. If vibration causes the meter pointer to move, the meter coil develops a back voltage, which dampens its movement.

The receive level position of S2 connects a circuit very similar to that already discussed for the send level position. In this position, the output of the receive level attenuator (passing through the input 6-db pad) is applied from R2 to contact 9 of section 3 front. Since contacts 9 and 6 are shorted in this position, the input signal is passed from contact 3, through a section of the bridge rectifier, CR5, contacts 5 and 12 of section 2 rear, to the negative side of M1. The d-c current (rectified by CR5) leaves the positive side of the meter and passes through contacts 12 and 5, another section of CR5 to ground, to return to E1. In this condition, the receive level attenuator may be adjusted until the meter reads 0 dbm.

In the PLATE CURRENT position of S2, switch contacts 12 and 6 of section 2 rear and 12 and 6 of section 1 rear connect the meter, M1, across the d-c amplifier metering resistors, R30 and R31. Since the current through these resistors are in opposite directions, the voltage applied to the meter is equal to the difference of the potentials across R30 and R31.

In effect, the meter reads the differential plate current. The differential current for mark signals cause the meter to read to the right, and space signals cause a deflection to the left. The current reading of the meter represents the current effective in producing the magnetic flux in the receive relay, as discussed.

The meter connection in the RECEIVE BIAS position is similar to that already discussed for the send bias position. R65 again applies a positive potential to the negative terminal of M1. R65 is not paralleled by R69 as was the case in the SEND BIAS position of S2. The positive terminal of the meter connects through contacts 12 and 7 of section 1 rear, through R66 to C23. R67 is now connected through contacts 1 and 2 of section 3 rear to the armature 6 of the receive relay, K1. When the armature is in the MARK position, a positive voltage from the +105-volt supply charges C23 through R67, R73, contacts 1 and 2 of section 4 rear, and R68. When the receive bias control is properly adjusted, the positive potential on both sides of the meter will be equal, permitting it to indicate a zero reading.

QUIZ

1. The limiter-amplifier (fig. 11-2) applies a constant voltage to the input circuit of the discriminator, and variations at its input may swing between what limits?
2. The bandpass filter, Z1 (fig. 11-3) is used for the reception of wide frequency-shift signals, and has an input impedance of what value?
3. What is the purpose of the axis restorer circuit (fig. 11-3)?
4. Tuning of the oscillator (fig. 11-3) is accomplished through the use of what control?
5. The output of the tone modulator (fig. 11-3) is keyed from "tone-on" to "tone-off" by a high negative bias voltage applied to its grids from what circuit?
6. The electronic relay (fig. 11-3) employs tubes V14 and V15, which perform what function?
7. Two double diodes V16 and V17 (fig. 11-5) are connected together in a special circuit for what purpose?
8. Conversion of the current pulses to audio tones is accomplished by an audio oscillator in the tone converter (fig. 11-7), which operates at what frequencies?
9. Two germanium rectifiers at the output of the frequency discriminator (fig. 11-7) rectify and convert the respective frequencies to what type signals?

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10. The cathodes of the d-c amplifier pentodes (fig. 11-8) are connected to opposite ends of what control?

11. If the received marking impulses are longer than normal, the condition is known as what type bias?

12. (a) The current in the 8-1 winding of K2 (fig. 11-8) is what value during mark?
(b) What is the value of this current during space?

13. With the control switch, S3 (fig. 11-8) in the AUTO position, an incoming signal will have what effect on the tone terminal?

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CHAPTER 12

COMMON MAINTENANCE PROBLEMS: RADIO AND RADIAC

THE NAVY MAINTENANCE PROGRAM

The efficient operation of all electronic equipment requires expert maintenance, both preventive and corrective. Further, the heavy workload imposed upon naval electronic equipments, and the complicated and varied interconnections of the many units in the various systems necessitate the adoption and use of an effective corrective and preventive maintenance program. Preventive maintenance procedures should be well organized to include all parts of the equipment itself, and those units or materials upon which the operation of the equipment depends. Corrective maintenance requires a working knowledge of the unit to be repaired, and should be built around well organized procedures and properly executed actions.

This chapter is primarily concerned with preventive maintenance, using the POMSEE program as its structural basis. It also includes many common maintenance problems of the AN/SRT-14, 15, and 16 transmitters and the AN/SRR-11, 12, and 13 receivers, as indicated in failure reports sent to the Bureau of Ships over a considerable period of time. The discussion covers both the preventive and corrective aspects concerning these failures. For a detailed outline of the entire maintenance procedures for the equipments discussed in this chapter, refer to the instruction book for that particular unit.

PREVENTIVE MAINTENANCE

Many ships and naval stations were able to develop their own preventive maintenance program, which was adequate to meet their particular operating needs. On the other hand, many ships and stations were unable to cope with the maintenance problem, which in some cases resulted in failing to accomplish their assigned naval mission. An efficient preventive maintenance program was found to improve equipment operation, and to decrease the casualty and repair tasks resulting in long and tedious hours of corrective maintenance. Consequently, the haphazard programs employed by some ships and stations resulted in the establishment of a standardized preventive maintenance program (POMSEE) by the Bureau of Ships.

Preventive maintenance is the systematic accomplishment of items deemed necessary to reduce or eliminate failures and prolong the useful life of the equipment. This involves routine checking of serviceable equipment to avoid needless breakdown, and keeping the equipment operating at its maximum efficiency. If wear on a particular item or a basic malfunction is detected and corrected in its early stages, a serious shutdown may be prevented.

To effect preventive maintenance, accurate records must be kept regarding routine checks and periodic maintenance procedures conducted, and those that are required. An effective approach to maintenance is the performance, operational, and maintenance standards for electronic equipment (POMSEE) program.

POMSEE

The POMSEE program serves as the basis for the recommended preventive maintenance program for all shipboard electronic equipment (also shore-based installed equipment) under the technical control of the Bureau of Ships.

Basically, the procedures for the operator and technician are outlined in three publications to be used for guidance in performing checks, comparing standards, and for the proper operation of each particular equipment for which the books, sheets, and/or charts are issued.

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The expression, POMSEE, means PERFORMANCE, OPERATIONAL, AND MAINTENANCE STANDARDS FOR ELECTRONIC EQUIPMENTS. The Bureau of Ships has supplemented the existing instruction books or technical manuals with two separate publications, which make up the maintenance section of the POMSEE program and one chart, which summarizes the proper operating procedures of a particular equipment.

Performance Standards Sheets

Performance Standards Sheets (fig. 12-1) provide the operational and technical data indicative of the minimum acceptable level of performance for electronic equipment. A binder entitled, *Binder for Electronic Equipment Performance Standards Sheets,* NavShips 93,000, is provided for incorporating all sheets under one cover. This binder is classified CONFIDENTIAL even though many of the performance standards sheets contained therein will be unclassified. This has been allowed so that sheets for all ships or stations installations may be filed in a single record.

The short title (ordering number) number for each sheet will consist of the NavShips number of the basic technical manual, plus the decimal number, .32 (for example, 91921(A).32).

In some instances, the above rule does not apply, because POMSEE publications often cover several equipments, and each equipment may have its own technical manual with a different NavShips number. Incorrect requisitions clutter up the supply system and slow down the processing of correct requisitions. To expedite requests for POMSEE publications, DO NOT requisition any electronics publications from the supply system until availability notices and NavShips numbers have appeared in the EIB or NavSandA 2002.

The Electronic Equipment Performance Standards Sheets (short title 9xxxx.32) assembled in the *Binder For Electronic Equipment Performance Standards Sheets,* NavShips 93,000, provide for each equipment type a single standard, which furnishes both a nontechnical and technical description of the expected equipment performance. This standard will be met by all ship installations of the particular model and will ensure optimum performance under average conditions of propagation.

Performance Standards Sheets will be distributed directly to ships and stations based on the inventory reported on NavShips 4110 at that time. It is intended that each ship and station will have only those Performance Standards Sheets applicable to their installed electronic equipment.

The Performance Standards Sheets are made up of two parts: Part I, Operational Performance and Part II, Standards for Equipment Measurement. These parts are self-explanatory. Availability of each new sheet will be announced in the EIB when distribution has been made.

Maintenance Standards Books

Maintenance Standards Books provide standard methods for determining measurements affecting the performance of a specific equipment, space to record such measurements, and a preventive maintenance schedule for the equipment. The Maintenance Standards Book is divided into two parts, which includes Part I, Test Procedures and Maintenance References (formerly Performance Standards Book); and Part II, Preventive Maintenance Check-off (formerly Maintenance Check-off Book). Information regarding the availability of these books is published in the EIB. The short title number for each book will consist of the NavShips number of the basic technical manual plus the decimal number .42 (i.e. 91921(A) .42, formerly .31 and .41). Use caution in requisitioning these books so as not to clutter the supply system.

Holders of the older POMSEE publications are instructed, according to the Bureau of Ships Instruction, 9670.86A, to change the title of all Performance Standards Books to Maintenance Standards Books, Part I, Test Procedures and Maintenance References, and to change the title of all Maintenance Check-off Books to Maintenance Standards Books, Part II, Preventive Maintenance Check-off.

The *Maintenance Standards Book,* NavShips 9xxxx.42, is composed of the following three basic sections, and all Maintenance Standards Books will contain these sections within their covers. However, some of these sections may contain several pages, and/or they may be alphabetically or numerically shifted to suit the programming and production methods of the manufacturer.

The first section, front matter, in all maintenance books will contain the following information pages in the order listed below, except where the numerical order of specific pages

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PERFORMANCE STANDARDS SHEET FOR
RADIO RECEIVING SETS AN/SRR-11, AN/SRR-12, AN/SRR-13A

TABLE I - OPERATIONAL PERFORMANCE
The many variables involved, such as radio propagation conditions at the time of test, power output of the transmitter being received and the type of antenna installation, preclude definitive predictions of operating ranges. To achieve optimum reliable ranges, frequency selection should be made in accordance with DNC-14, as corrected by half-hourly radio propagation predictions given over WWV and WWVH.

TABLE II - STANDARDS FOR EQUIPMENT MEASUREMENT

MEASUREMENT	SECTION AND STEP	STANDARD
Bandwidth (AN/SRR-11)	Section F Step 3	2 kc min. (at 6 db)
Bandwidth (AN/SRR-12)	Section F Step 3	4 kc min. (at 6 db)
Bandwidth (AN/SRR-13A)	Section F Step 3	1 kc min. (at 6 db)
Receiver Sensitivity (Sr)	Section F Step 1.1	7 μV max
Performance Figure (PF)*	Calculate	7 max

* PF = Sr (μV)

Note:

1. Total time required to accomplish measurements of Table II - 1 hour.
2. All sections and steps refer to the tests in Part I of the Maintenance Standards Book, NAVSHIPS 91875.42.

Figure 12-1. –Performance Standards Sheet.

may occasionally be reversed, or one or more specific topics may be omitted.

1. List of Effective Pages. This page must always be carefully checked to ensure that all applicable pages are contained therein.
2. Promulgating Letter.
3. Record of Entries and Corrections. This page is used to record the lists in item one.
4. Record of Field Changes. This page is very important in that repairs affected

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to an equipment must include all circuitry to be useful and effective.

5. Introduction. These pages contain a dedetailed explanation of the contents contained therein and directions on how to use the publications.
6. Test Equipment. Here is included all test equipment needed to perform the tests indicated.
7. Special Procedures. These pages specify any condition peculiar to a certain test and should be followed to the letter.
8. Block Diagram (fig. 12-2). In most cases this will be a simplified block diagram to provide an over-all picture of the entire system. It can be very helpful in locating a particular chassis section or stage of operation. Sometimes this section will be included as page 1-0, Part I, Test Procedures and Maintenance References, rather than in the front matter.

The second section, Test Procedures and Maintenance References, Part I, is composed of several subsections. Each subsection contains the procedures for obtaining the maintenance standard test indications, which are given in a series of charts, each chart (fig. 12-3) or group of charts covering a functional section of the entire system. An illustration page (fig. 12-4) facing each chart page shows the equipment setup pertaining to each of the procedural steps of that chart page, and each setup carries a step number (enclosed in a circle) corresponding to the step of the chart to which it applies. On the chart page the section letter at the top of the page indicates the section to which the chart procedure refers (fig. 12-2). The step numbers indicating the procedure to be followed are enclosed in the circles, and correspond with the numbers in the illustration of figure 12-3. This illustration shows the equipment setup pertaining to each of the procedural steps of that chart page. Arrows extending from the step number present certain basic information in more detail such as the point of connection of a switch or control, or the indicator from which the test reading is to be taken. Each subsection is designated by a letter (A, B, C, etc). These subsections are further identified on the block diagram (fig. 12-2).

Part I consists of a series of tests that, when completed, will indicate the relative performance of the equipment. These tests and measurements, known as reference standards or performance standards, are made at critical or

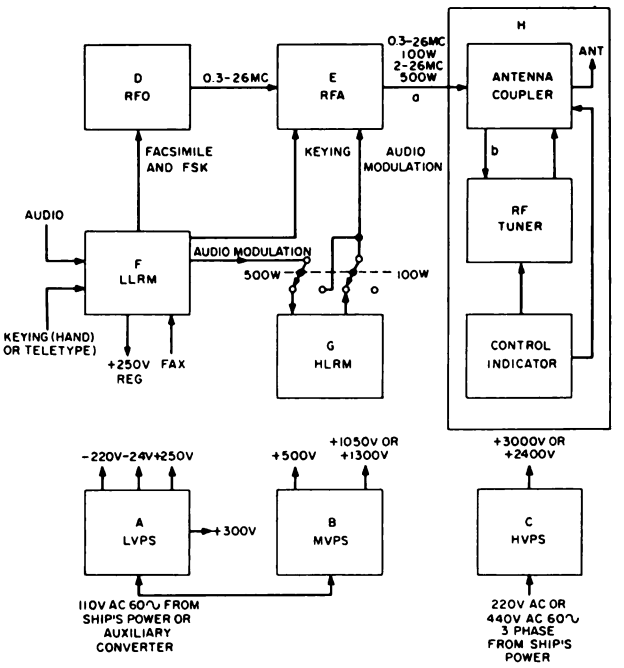

Figure 12-2. —Block diagram and lettering of the transmitter major units.

significant points when the equipment is known to be performing at the maximum of its capabilities. REMEMBER that these reference standards (Performance Standards) apply only to the equipment to which this particular book has been permanently assigned, and because of this individuality they are of great value only to that equipment.

Upon receipt of the Maintenance Standards Book the standards are to be established and the Reference Standards Summary Sheet (Performance Standards Summary Sheet) discussed later are to be completed and forwarded to the Bureau. These standards should be reestablished only after each overhaul or a major field change has been accomplished. Prior to this establishment, each section must be checked thoroughly to ensure that the equipment is operating at its maximum capability. After these checks have been accomplished, the test and measurement results should be entered, in ink, in the performance standard or reference standard column. Care must be exercised at all times when making these measurements to ensure that the correct procedures are implicitly followed otherwise the recorded standards will become meaningless.

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Steps 1 thru 5

LOW VOLTAGE POWER SUPPLY

Operating Conditions and Control Settings:

Transmitter adjusted for 100-watt, CW operation.

Test Equipment Required:

Multimeter AN/PSM-4 or ME-25A/U

STEP NO.	STEP ACTION REQUIRED	PRELIMINARY ACTION	READ INDICATION ON	PERF. STD.
1	Record primary supply voltage.	Connect Multimeter AN/PSM-4, using 250-volt, AC scale, between F3007 and F3008.	Multimeter AN/PSM-4	_____ VAC (100 to 120)
2	Record output voltage of 300-volt power supply.	Set VOLTMETER switch (S1384): LV. Record indication with transmitter adjusted for each of the frequencies indicated in PERF. STD. column.	VOLTMETER (M1301)	mc .35 ___ VDC 2.5 ___ VDC 5.5 ___ VDC 6.5 ___ VDC 11.0 ___ VDC 15.5 ___ VDC 16.5 ___ VDC 21.0 ___ VDC 25.5 ___ VDC (270 to 330)
3	Record output voltage of —220-volt power supply.	Set VOLTMETER switch (S1384): BIAS. Record indication with transmitter adjusted for each of the frequencies indicated in PERF. STD. column.	VOLTMETER (M1301)	mc .35 ___ VDC 2.5 ___ VDC 5.5 ___ VDC 6.5 ___ VDC 11.0 ___ VDC 15.5 ___ VDC 16.5 ___ VDC 21.0 ___ VDC 25.5 ___ VDC (—210 to —230)
4	Record output voltage of 250-volt power supply.	Connect Multimeter AN/PSM-4, using 500-volt, DC DIRECT scale, J3005 to ground.	Multimeter AN/PSM-4	_____ VDC (240 to 260)
5	Record output voltage of 250-volt, regulated power supply.	Connect Multimeter AN/PSM-4, using 500-volt, DC DIRECT scale, from J1110 to ground.	Multimeter AN/PSM-4	_____ VDC (250) (see page ii.)

Figure 12-3. –Sample chart for obtaining performance standards.

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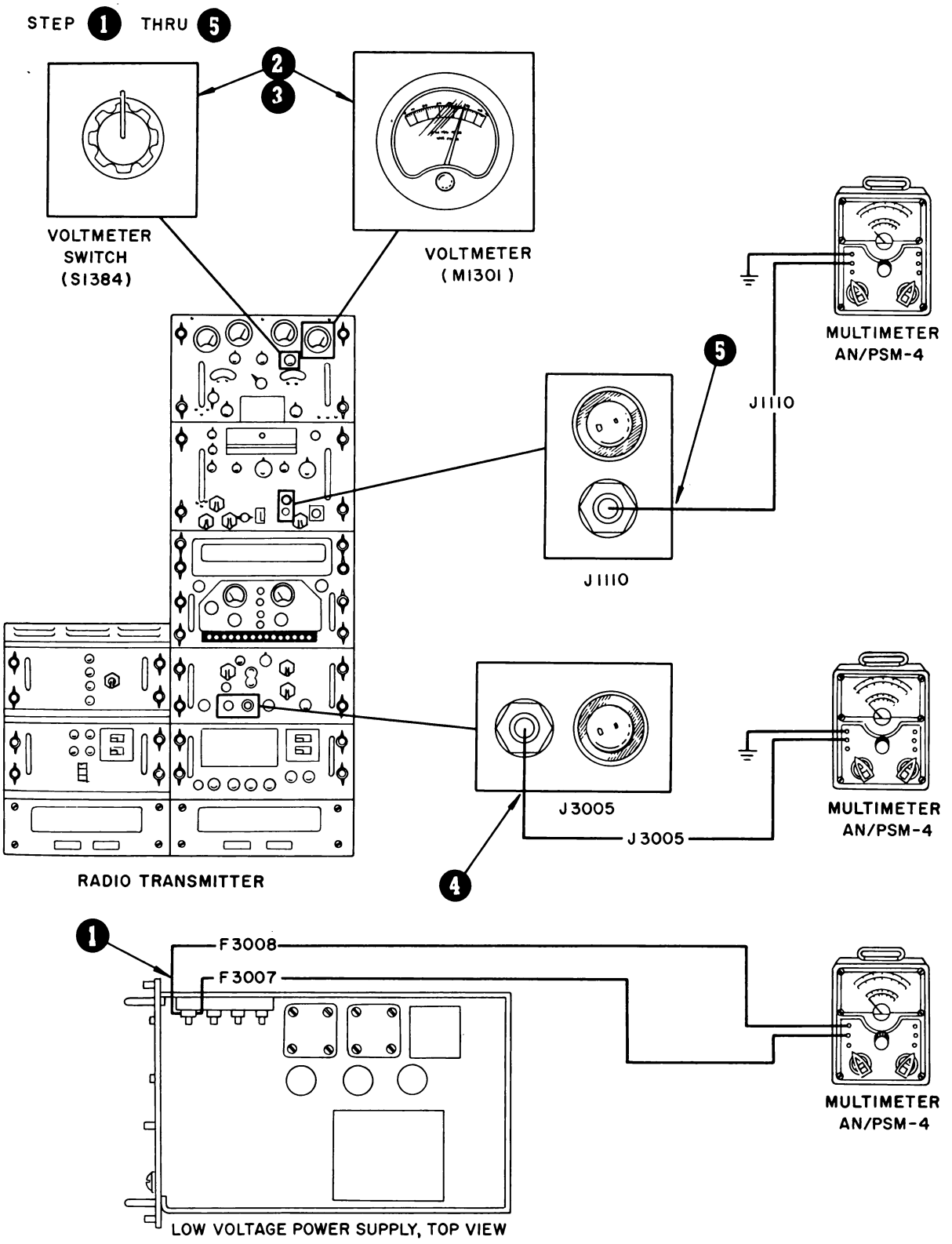

Figure 12-4. –Sample diagram of step procedures.

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When the equipment performance drops below the minimum acceptable standard, the appropriate instruction book should be consulted for more detailed service and/or repair procedures.

In some cases the illustrations are used for the reference standards and/or preventive maintenance sections. On those illustrations used for both reference standards and preventive maintenance, the steps are designated in some maintenance books by the letters (O.M.) in the step column. Whereas, in other books, the preventive maintenance steps are denoted by a white circle with black figures, and the reference standards are denoted by a black circle with white figures. The procedures designated by step numbers enclosed in stars are referred to in the Performance Standards Sheet (9xxxx.32) for that equipment.

REMEMBER THAT THE MAINTENANCE STANDARDS BOOK APPLIES ONLY TO THE EQUIPMENT FOR WHICH THE BOOK HAS BEEN PERMANENTLY ASSIGNED.

The last section, Preventive Maintenance Check-off, Part II, contains a series of tests, which provide a systematic and efficient method of checking equipment, and of performing preventive maintenance.

Charts are provided for the initials of the person performing the checks (technician and/or operator). If the result of the check is a measurable quantity, space is provided for recording the result.

The charts in this section show the check procedures to be performed by the maintenance technician and/or operators. The check procedures are schedules for regular periods (daily, weekly, etc).

A complete list of the check periods that may be contained in any Maintenance Check-off Book includes daily (fig. 12-5), weekly, monthly, quarterly (fig. 12-6), semiannually, nonperiodic, and annual period pages. All check periods for the various types of equipments will not be the same.

Upon completing each check prescribed in the charts, enter the results in the time-schedule tables accompanying the charts. These entries are of prime importance, because they will indicate whether or not the equipment is performing properly. Comparison of a reading with readings previously obtained, along with the initial maintenance standard test indications, will quickly reveal any significant change. Slight changes are expected from time to time and need not cause alarm. However, when a particular step shows a reading that varies progressively in the same direction, it is an indication of improper operation or of impending failure, and corrective measures should be taken.

Some Maintenance Check-off Books contain additional features, which are very helpful for those responsible for performing the maintenance check. At the top of each page is a list of operating conditions and control settings (fig. 12-6) for that particular page. Following this, on the chart page for each period, is a list of test equipment, if any, required to perform the procedures within that period.

Operating Instruction Charts

Operating instruction charts provide a summary of the procedures to be followed in starting, operating, and stopping the equipment. A chart is provided for installation at each operating position. The short title number for each chart will consist of the NavShips number of the basic technical manual and the decimal number (.21) except for those charts provided via the supply officer (Code 516), Philadelphia Naval Shipyard.

Performance Standards Summary Sheets

The first page of all Maintenance Standards Books is an insert-type page for recording the Performance Standards Summary (fig. 12-7), also known as the Reference Standards Summary (fig. 12-8). After the equipment has been brought up to optimum performance, the standards accomplished, and all test indications have been entered in the Maintenance Standards Book, the Performance Standards Summary (Reference Standards Summary) sheets are completed from the records obtained in the procedure. These sheets are then forwarded to the Chief, Bureau of Ships, Navy Department, Washington 25, D.C., Attn: Code 965.

Note that the section letter and step number permit identifying where the readings and measurements as standards are recorded.

In-port Procedures

In general, in-port procedures will not require the the equipment to be energized daily for the sole purpose of making the daily checks. However, the equipment should be energized at least twice a week for checking and at least two

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Part II — Quarterly

Steps 1 and 1

AN/URR-27

Operating Conditions and Control Settings:

As given in Special Procedures.

Test Equipment Required:

Audio Oscillator TS-382/U
Multimeter AN/USM-34
R. F. Signal Generator Set AN/URM-25
R. F. Signal Generator Set AN/URM-26

STEP NO.	STEP ACTION REQUIRED	PROCEDURE
1	Record audio gain.	Perform step 1 of Section B, Part I. Obtain correct indication and record.
1	Record audio response at high and low ends of range.	Perform step 2 of Section B, Part I. Obtain correct indications and record.

Time Schedule: Record and initial. Approx Time Reqd for Quarterly Checks — 1 hr

	1st Year				2nd Year			
Quarter	___Quarter 19___	___Quarter 19___	___Quarter 19___	___Quarter 19___	___Quarter 19___	___Quarter 19___	___Quarter 19___ *	___Quarter 19___
Step 1								
Step 2 L								
Step 2 H								
Initial								

Figure 12-5. –Chart page used with the POMSEE program.

days before getting underway. Enter IN-PORT in the appropriate blanks.

Summary

In general, the Maintenance Standards Book, Part I, Test Procedures and Maintenance References, provides an itemized step-by-step procedure, which enables the engineer or other person making the standard tests to note critical or significant operating values (meter and dial readings, etc) representative of optimum operating conditions. Upper and lower limits or tolerances for dial readings, voltages, or currents are given so that an indication is readily available if performance is below the allowable limits. However, no attempt is made in the Maintenance Standards Book to show how to locate a trouble, but by comparing the readings with established critical circuit readings the ET may isolate the trouble.

Reference to the instruction book for the particular equipment is still required for effective troubleshooting or corrective maintenance.

The Maintenance Standards Book, Part II, Preventive Maintenance Check-off, provides spaces for recording the maintenance checks performed over a two-year period, requires that standard tests be performed at regular intervals on circuits and components, and specifies what and when other routine maintenance such as lubrication is to be accomplished. By the proper use of this book, routine checks and preventive maintenance may be accomplished in a systematic manner.

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Step 1

Operating Conditions and Control Settings:

None

Test Equipment Required:

None

STEP NO.	STEP ACTION REQUIRED	PRELIMINARY ACTION	READ INDICATION ON	PERF. STD.
1 **	Perform operational test on transmitter.	Operate transmitter from shut-down condition, observing the operation of all controls and indicator lights. When all indications and functions are normal, initial appropriate space below.	Indicator lights and panel meters	

Time Schedule: Initial.

Approx Time Reqd for Daily Step — 15 min

Day	Step	Jan. 19__	Feb. 19__	Mar. 19__	Apr. 19__	May 19__	June 19__	July 19__	Aug. 19__	Sept. 19__	Oct. 19__	Nov. 19__	Dec. 19__
1	1												
2													
3													
4													
5													
6													
7													
8													
9													
10													
11													
12													
13													
14													
15													
16													
17													
18													
19													
20													
21													
22													
23													
24													
25													
26													
27													
28													
29													
30													
31													
Initial													

Figure 12-6. —Preventive Maintenance Check-off, Part II, Quarterly Sheet.

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RADIO TRANSMITTING SET AN/SRT-14, -15, -16

PERFORMANCE STANDARD SUMMARY

Model _____

Serial No. _____

Installed In _____
Ship or Station

After the equipment has been brought up to optimum performance and the performance standards accomplished, record on this Summary-sheet the Standards which have been entered in the book. Forward this Summary-sheet to Chief, Bureau of Ships, Navy Department, Washington 25, D. C., Attn: Code 965.

Section and Step	Perf. Std.	Section and Step	Perf. Std.	Section and Step	Perf. Std.	Section and Step	Perf. Std.	Section and Step	Perf. Std.
A1	_____ VAC		_____ VDC		VDC	D1	_____ cps		_____ ma
A2	_____ VDC		_____ VDC		VDC	D2	_____ V		_____ ma
	_____ VDC		_____ VDC	C1	VAC		_____ c		_____ ma
	_____ VDC		_____ VDC		VAC		_____ V		_____ ma
	_____ VDC		_____ VDC		VAC		_____ c		_____ ma
	_____ VDC		_____ VDC	C2	VDC	D3	_____ VAC		_____ ma
	_____ VDC		_____ VDC		VDC	D4	_____ VAC		_____ ma
	_____ VDC	B2	_____ VDC		VDC	D5	_____ VDC		_____ ma
	_____ VDC		_____ VDC		VDC	D6	_____ Ratio	E3	_____ ma
	_____ VDC		_____ VDC		VDC	D7	_____ Ratio		_____ ma
A3	_____ VDC		_____ VDC		VDC	D8	_____ Ratio		_____ ma
	_____ VDC		_____ VDC		VDC	D9	_____ Ratio		_____ ma
	_____ VDC		_____ VDC		VDC	D10	_____ Ratio		_____ ma
	_____ VDC		_____ VDC		VDC	E1	_____ ma		_____ ma
	_____ VDC		_____ VDC	C3	VDC		_____ ma		_____ ma
	_____ VDC		_____ VDC		VDC		_____ ma		_____ ma
	_____ VDC	B3	_____ VDC		VDC		_____ ma		_____ ma
	_____ VDC		_____ VDC		VDC		_____ ma	E4	_____ ma
	_____ VDC		_____ VDC		VDC		_____ ma		_____ ma
A4	_____ VDC		_____ VDC		VDC		_____ ma		_____ ma
A5	_____ VDC		_____ VDC		VDC		_____ ma		_____ ma
B1	_____ VDC		_____ VDC		VDC		_____ ma		_____ ma
	_____ VDC		_____ VDC		VDC	E2	_____ ma		_____ ma

Figure 12-7. —Reference Standard Summary Sheet.

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REFERENCE STANDARDS SUMMARY

Model _____

Serial No. _____

Installed In _____
(Ship or Station)

After the equipment has been brought up to optimum performance, and standards accomplished, record on this Summary sheet the test indications which have been entered in this book. Forward this Summary sheet to Chief, Bureau of Ships, Navy Department, Washington 25, D. C., Attn: Code 975.

Step No.	Ref. Std.
Section A	
1	_____ VAC
2	_____ VAC
3	_____ div
4	_____ div
Section B	
1	_____ cps
	_____ cps
2	_____ cps
	_____ cps
3	_____ VAC
	_____ VAC

Step No.	Ref. Std.
Section C	
1	_____ μv
	_____ μv
	_____ μv
2	_____ kc
3	_____ VAC
	_____ VAC
	_____ VAC
	_____ VAC
	_____ VAC

Step No.	Ref. Std.
	_____ VAC
	_____ VAC
	_____ VAC
4	_____ μv
5	_____ VDC
6	_____ db
Section D	
1	_____ div
2	_____ div
3	_____ div
Section E	
1	_____ ohm
2	_____ megohms

List all field changes which have been accomplished on this equipment _____

_____

Signature _____

Title-Position _____

Date _____

Figure 12-8. –Performance standards summary sheet.

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Generally, the same step required for determining Performance Standards must be repeated by the ship's force in making routine checks for entry in the Maintenance Check-off Books. Many of the daily, weekly, and monthly checks prescribed should be done by operators, as outlined in the Check-off Books. However, the individual ship's EMO will determine who should undertake each equipment under test.

Always REMEMBER:

1. That the reference column tolerances shown (in parentheses) in Part I are not absolute limits.
2. That step numbers enclosed in stars are referred to in the Performance Standards Sheets.
3. To record all tests on the Summary Sheet along with a list of all Field Changes that have been made, and forward the sheet to the Bureau of Ships.
4. To enter the correct date at the top of the Checks Column in Part II. Appropriate dates should be entered at the top of subsequent columns when initiated at a later date.
5. To RECORD THE SERIAL NUMBER OF THE EQUIPMENT TO WHICH THE BOOK IS ASSIGNED, IN INK, IN THE SPACE PROVIDED ON BOTH THE COVER AND THE TITLE PAGE.

THE TEST STANDARDS MEAN NOTHING UNLESS THE INITIAL MAINTENANCE STANDARD TEST AND SUBSEQUENT CHECKS ARE COMPARED AGAINST THEMSELVES.

COMMON TRANSMITTER MAINTENANCE PROBLEMS

PREVENTIVE MAINTENANCE

Radio Transmitters AN/SRT-14, 15, and 16, are composed of six assemblies: radio frequency oscillator (RFO), radio frequency amplifiers (RFA), antenna tuning equipment (antenna coupler and r-f tuner), low level radio modulator (LLRM), high level radio modulator (HLRM), and power supplies (100 and 500 watt). Most of these assemblies can be removed and replaced by loosening a few captive screws. The major assemblies are divided into subassemblies, many of which are easily connected or disconnected from the major assemblies. For example, in the RFO the crystal oscillator subassembly (unit 1) may be easily removed by disconnecting two 1/2-turn cowl fasteners and removing the five cables connected to the main chassis.

All assemblies and subassemblies of the AN/SRT-14, 15, and 16 are interchangeable. In the naval installation where there are several AN/SRT-14's 15's, and 16's, the technician can maintain a high operating efficiency by making proper assembly and subassembly interchanges. In this way a defective drawer in a normally operative transmitter can be interchanged for a serviceable drawer in an already defective transmitter. The technician can then concern himself with the repair of several troubles in one transmitter rather than one trouble in several different transmitters. The power supply assemblies use conventional type tubes and transformers, and do not contain subassemblies.

One advantage of being able to interchange the assemblies and subassemblies can be appreciated in view of an operation where the technician needs every available transmitter to handle the work load and hasn't the time to repair each failure as it occurs.

After the transmitter has been installed, a thorough operational check will be performed by authorized personnel. At this time, the Maintenance Standards Book (with sheets similar to that of figure 12-3) will be completed. The optimum performance of the equipment will be such as to conform to the standards in the Performance Standards Sheet for that equipment.

Radio Frequency Oscillator

It has been found that frequency drift by either unit 3 (interpolation oscillator), unit 6 (10-kc step generator), or unit 8 (100-kc step generator) may cause a large error in the transmitter frequency output. For this reason, particular emphasis should be placed on the preventive maintenance procedure for these units, as outlined in the Maintenance Standards Book for the AN/SRT-14, 15, and 16 (page 31). It is further suggested that the outlined preventive maintenance procedure be conducted at least once weekly, or as often as required through your experience with the equipment. The corrective procedures are outlined later in this chapter.

Low Level Radio Modulator

Keying signals passing through the keying relay of the LLRM may cause burning or pitting

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of the relay contacts (see chapter 7, fig. 7-2 of this training course). If left unattended the condition will continue to develop until proper keying of the transmitter is impossible. This condition should be checked during the preventive maintenance procedure as often as is required.

C38 (fig. 7-2) in the +250-volt power supply of the LLRM is an 80-μf electrolytic capacitor. This capacitor alone makes up the filter capacitance for the power supply.

When replacement of C38 becomes necessary, care must be taken to procure a relatively new capacitor, or breakdown of C38 will recur in less than the usual time. If the replacement capacitors presently in stock are more than two years old (date stamped on case) they should be returned for newer ones. If the capacitors are over six months old, they must be reformed as outlined in EIB 475 of 6 Jan 1958.

Radio Frequency Amplifier

All meter readings should be taken under the same conditions and with the same antenna or dummy load. Weekly readings are to be made and compared with those of the previous week, so as to check the variations in the transmitter operating conditions.

Remember that various production line tubes will vary considerably in their current output capabilities. A definite trend or fluctuation in readings is to be interpreted as an indication of trouble. By charting these fluctuations, a reliable record of tubes and general transmitter performance is available for ready reference, and trouble may be avoided by changing tubes or correcting a fault before a critical stage is reached.

Antenna Tuning Equipment

The antenna coupler contains two actuator motors. One controls the loading switch, and the other controls the antenna transfer switch. These motors require lubrication and should be inspected, and cleaned by swishing in dry cleaning solvent. A small brush should be used to clean out any stubborn dirt or grease, which might tend to cling to the bearings. The bearings should be dried by spinning them in the air.

Specific lubrication periods in all cases should not be taken to be inflexible. For extreme operating conditions, these periods may possibly be reduced to a fraction of the originally specified time. Some abnormal conditions encountered may be extremely wide ranges of temperature, prolonged high speed operation, immersion in water, operation in dust or sand, or excessive exposure to moisture. Contamination of the lubricant will definitely take place under any of these conditions, and its protective quality will be lost.

It has been found through experimentation that when tuning the transmitter to any frequency within its range, care should be taken to maintain the reading of the position meter (discussed later) at a value above 50 μa. The position meter indicates the position of the sliding short on the main tuning coil in the r-f tuner. If the sliding arm of the coil is placed too near the ground potential, a serious mismatch will exist, and high standing waves will occur along the line. The standing wave lobes may be sufficient in amplitude to cause arcing between the center and grounded conductors of the coaxial line. When this occurs, all efforts to load the transmitter at any given frequency will fail.

It has been found that a particular frequency to which the transmitter is to be tuned will always occur at approximately the same setting on the position meter. Therefore, to minimize the occurrence of arc-over in the cable, the Electronic Technician should record the approximate reading of the position meter for various frequencies obtained successfully. The chart should be kept available for quick reference each time a change of frequency is required.

CORRECTIVE MAINTENANCE

The Electronic Technician must have a good working knowledge of a particular equipment before corrective maintenance can be adequately and safely performed. The circuit operation of the AN/SRT-14, 15, and 16 has been presented in chapters 5, 6, and 7 of this training course. A more detailed discussion of the equipment circuitry may be found in *The Instruction Book for the Radio Transmitting Sets, AN/SRT-14, AN/SRT-15, and AN/SRT-16,* NavShips 92121(A).

The test equipments required to effect corrective maintenance on these transmitters are outlined.

1. Multimeter ME-25A/U, 0 to 1000 v ac/dc in 7 ranges; 0 to 1000 ma in 6 ranges; 0 to 1000 megohms in 6 ranges.

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2. Oscilloscope OS-8A/U or Synchroscope AN/USM-32; sensitivity, 0.1 v (RMS) per inch; response, ± 3db from 30 cps to 2mc.
3. Frequency Meter AN/USM-29 or AN/FRM-3; 15 kc to 26 mc in 10-cps; accurate to within ± 1 cps in 1 mc.
4. Receiver Type RBA, RBB, RBC Series, or AN/SRR-11, AN/SRR-12, AN/SRR-13 series or equivalent; 15 kc to 500 kc, 0.5 to 4.0, mc 4.0 to 27.0 mc.
5. Audio Oscillator TS-382A/U; 20 to 200, 000 cps; ± 2 percent; output 0 to 100 milliwatts.
6. Dummy Load DA-91/U. 600 watt air cooled.
7. Signal Generator (RF) AN/URM-25; 10 kc to 50 mc; output 2 volts maximum.
8. Ammeter, d-c, Navy Type 60107 or Multimeter AN/PSM-4; 0 to 1000 microamperes in 5 ranges; accuracy ± 2 percent.
9. Resistance Bridge ZM-4/U; 1 ohm to 10 megohms; accuracy ± 0.15 percent.
10. Radio Frequency Bridge, Navy Type 60094; range 400 kc to 60 mc; resistance 0 to 1000 ohms ± 1 percent; reactance 0 to 5000 ohms at 1 mc ± 2 percent.
11. Variable Resistor; 1 megohm linear taper potentiometer.

It is not the purpose of this discussion to outline a detailed troubleshooting procedure for correcting all faults that may occur in the transmitter. Instead, those failures that have been found to recur frequently either because of weak components or improper operation, are emphasized.

Failure Reports

The use of the Electronic Failure Report Form DD787 to report each failure of the equipment is very important. Whether the failure is caused by a defective part, wear, improper operation, or an external cause, the breakdown should still be included in the report.

Each pad of forms includes full instructions for filling out the forms and forwarding them to the Bureau of Ships. The importance of providing complete information in the report cannot be too greatly emphasized. Most of the common equipment failures presented in this chapter were obtained from the failure report files of the Bureau of Ships.

Be sure you include the model number designation and serial number of the equipment (from the equipment nameplate), the type number and serial number of the major unit (from the major unit nameplate), and the type number and reference designation of the particular defective part (from the instruction book). Describe the cause of the failure completely, continuing on the back of the sheet, if necessary. Always keep a supply of failure report forms on board. You can procure these forms from the nearest Forms and Publications Supply Office.

Each failure report can help you and your division in several ways. It shows that you are doing your job, helps to insure the availability of replacement parts, helps to make your job easier, and gives you a chance to pass your knowledge to every man on the team.

Likewise, the failure report proves helpful to the Navy in evaluating its equipments, in improving future equipments, in maintaining and ordering replacement parts, in facilitating field changes, and in publishing necessary information concerning the operation and use of specific equipments.

Localizing the Trouble

The Electronic Technician should be familiar with the capabilities and limitations of the transmitter. Chapter's 5, 6, and 7 of this training course describe the function of the various components and circuits of a representative shipboard transmitter. Your ability to diagnose and localize troubles in the equipment will increase with experience and training.

By knowing the functions of the various major units of the transmitter, a malfunction can be quickly and easily localized. The functional block diagrams and schematic diagrams contained in the chapters mentioned above along with the instruction book, *Radio Transmitting Sets AN/SRT-14, 15, and 16,* NavShips 92121(A) should be used to examine the unit believed to be defective.

Each of the major units is contained in a separate drawer with facilities to extend the drawer from the cabinet and bypass the accompanying interlocks. Through a few simple operations, the drawer may be removed from the cabinet entirely. Always exercise caution when interlocks are bypassed.

If the trouble in the defective unit requires the replacement of a major component, considerable time may be consumed to bring the unit back to a serviceable condition. If another

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nonserviceable transmitter (AN/SRT-14, 15, or 16) is available, which contains a serviceable replacement drawer, substitute this drawer for the defective one while repairs are being made. In this way, the transmitter will be restored to service almost immediately.

Radio Frequency Oscillator

The radio frequency oscillator (fig. 12-9) is composed of 14 plug-in units and a chassis consisting of a mounting rack for the 14 subunits. A frequency is set up by rotating knobs to the desired frequency in conjunction with the table mounted on the RFO front panel door to determine settings of the 5 MC knob (AA) and the 1 MC knob (BB). Figure 12-10 shows a simplified block diagram of the radio frequency oscillator.

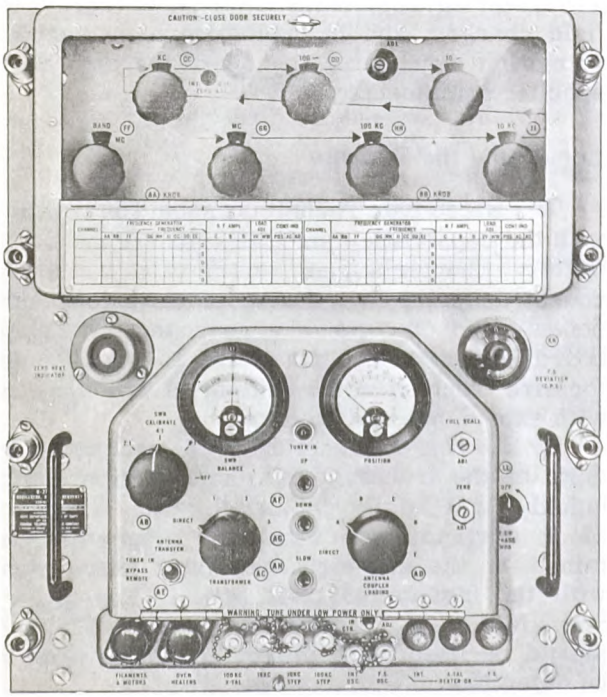

Figure 12-9. –Radio frequency oscillator front panel controls and indicators.

When a frequency is set up, but there is no output from the RFO, as indicated on the radio frequency amplifier voltmeter (with the voltmeter switch in the RF-IN position and the excitation control fully clockwise), one or more of the RFO subunits or the cabinet interconnecting cable may be faulty. If the RFO output fails during transmission, the carrier-on light will remain on, but the RFA panel-mounted voltmeter (with the voltmeter switch in the RF-IN position) will show zero input. The voltmeter should show an input of from 2 to 5 volts, depending on the setting of the excitation control (fig. 12-11).

To determine that the lack of RFO output is not due to the low level radio modulator keying circuit, check the change in the KEY V IN at the RFO test socket (fig. 12-12) as the transmitter is keyed on and off. With the test key in the center position (transmitter unkeyed) the multimeter should read approximately -30 volts. With the test key operated (transmitter keyed) the meter should read zero (0) volts.

After it has been determined that the LLRM keying circuit is operating properly and there is still no RFO output, the following checks should be performed:

1. Check units 11A, 11B, and 11C, (fig. 12-10) by setting up three different frequencies. The first frequency should be in the range from 0.3 to 6 mc, the second frequency between 6 and 16 mc, and the third within 16 to 26 mc. If only the second and third frequencies give an RFO output, as indicated on the RFA voltmeter (fig. 12-11) with the voltmeter switch in the RF IN (5 v) position, unit 11A is probably at fault. If only the first and third frequencies give an RFO output, unit 11B is probably at fault. If only the first and second frequencies give an RFO output, unit 11C is probably at fault.

2. Check for an RFO output during FSK and FAX operation. Switch the service selector on the low level radio modulator alternately in the FSK and FAX positions. If there is no RFO output during FSK operation, unit 12 (frequency shift oscillator) of the RFO is probably at fault. To check the frequency of any signal fed to the oscilloscope, Lissajous patterns should be used. Try to get the smallest frequency ratio possible, preferably a 1:1 ratio.

3. Check the 100-kc output of the crystal oscillator by connecting the oscilloscope to the 100 KC XTAL jack on the RFO front panel, or by means of the WWV frequency check standard. When using a single scope to make this check, only the presence of the unit 1 output can be determined. A two-to-one sinusoidal ratio obtained with the OS-8A/U (maximum sweep frequency 50 kc) or a one to one ratio obtained with the AN/USM-32 (maximum 200 kc sweep) should

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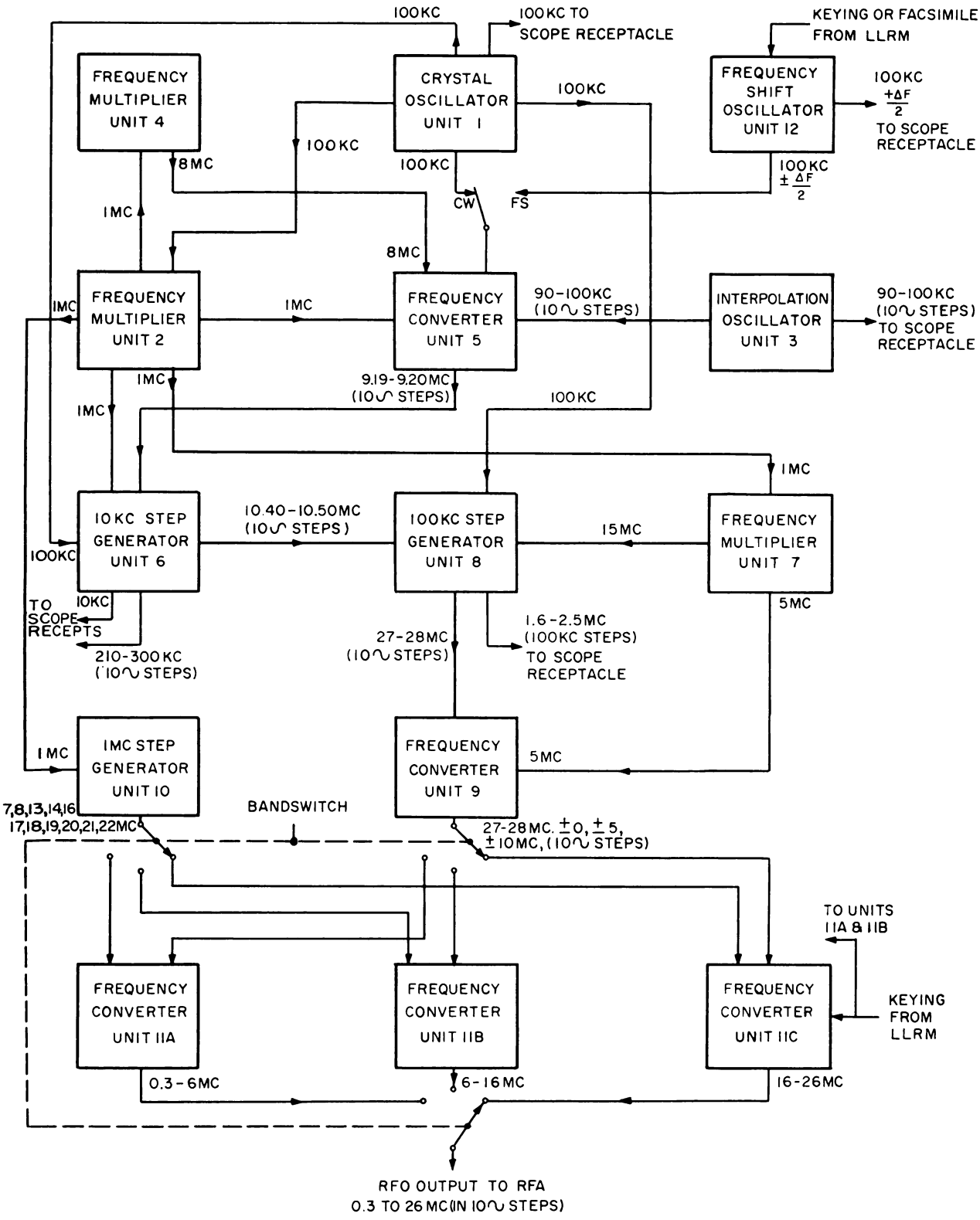

Figure 12-10. –Radio frequency oscillator block diagram.

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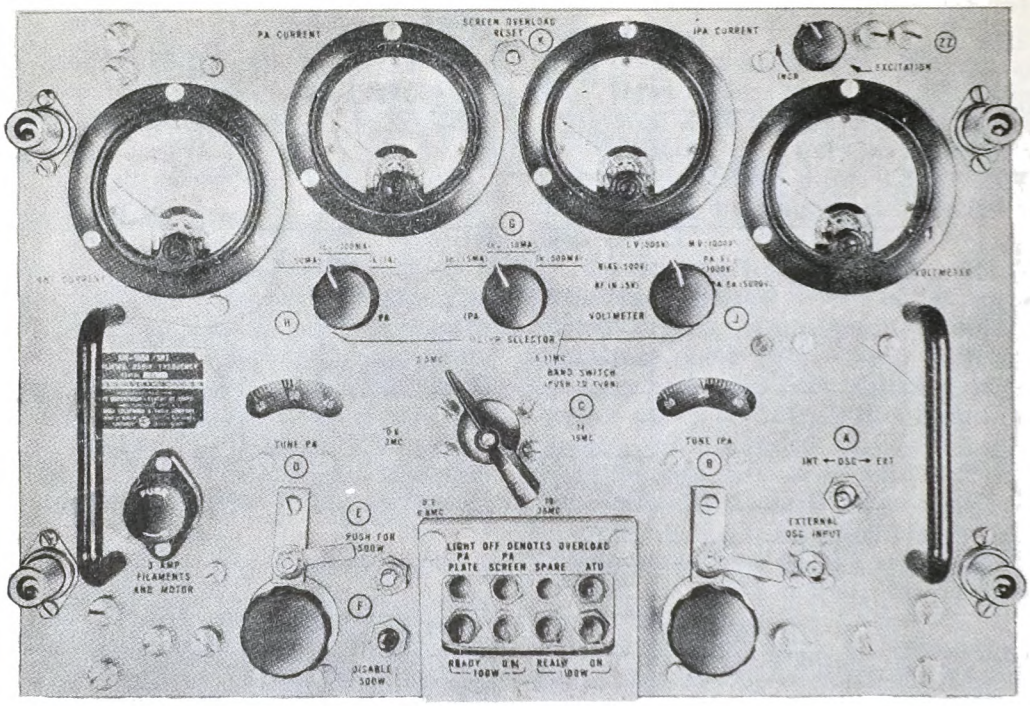

Figure 12-11. –Radio frequency amplifier front panel controls and indicators.

not be relied upon as an adequate frequency check for the unit 1 oscillator. The sweep circuits of these oscilloscopes are not designed to serve as frequency check standards.

If unit 1 of another transmitter group is known to be operating at the correct frequency through a previous check against station WWV, it may be used to check the frequency of the unit 1 oscillator in question. A Lissajous pattern may be obtained by connecting one of the 100 KC XTAL outputs to the X axis of the oscilloscope, and the other to the Y axis.

The frequency output of unit 1 can be changed by adjusting L1 (figure 6, Chapter 5). If a zero beat can not be obtained by adjusting the top slug of inductance L1, place unit 1 on a subunit extension test cable (if supplied) and adjust the bottom slug of the inductor. The bottom slug of L1 is factory adjusted, and should not be touched unless absolutely necessary.

4. Check the 90- to 100-kc output of unit 3 (interpolation oscillator in figure 12-10). Unit 3 is the most critical unit in the RFO. Extreme care should be taken to prevent changing the variable inductors during servicing. In the event that tests indicate that unit 3 is in need of alignment, always attempt to zero the unit with the INT OSC ZERO ADJ on the front panel. Before it can be determined that unit 3 is at fault, it must have already been determined that unit 1 is operating properly as discussed above.

Whenever the oscillator tube of unit 3 is replaced, it generally throws the unit out of alignment. This does not create a serious problem since the step accuracy of the 1,000-, 100-, and 10-cps switches is retained. To correct the

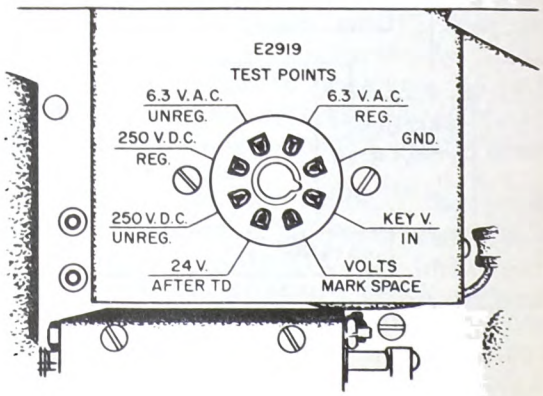

Figure 12-12. –Radio frequency oscillator metering socket.

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error, zero the INT OSC ZERO ADJ, which may be done without removing the RFO drawer.

The alignment of unit 3 is a critical process and should be undertaken only by the most skilled technical personnel. The circuits of the interpolation oscillator are of such precision and accuracy that alignment of its tuned circuits should not be attempted by field personnel. These procedures are to be accomplished through the use of tender or shipyard facilities.

5. Check the outputs of the 10-kc step generator, unit 6. Unit 6 provides three outputs, two of which can be checked with an external oscilloscope, from the RFO front panel. The 10-kc output of the blocking oscillator (see V4A, Figure 12, chapter 5) may be observed at the 10-kc jack on the RFO, (fig. 12-9) and the 210 to 300 kc output of unit 6 can be checked at the 10 kc-step jack. If either of these outputs is not available at its respective jack, check for a defective tube in the circuit indicated. To make a check of the third output, it is necessary to extend the RFO chassis from the cabinet, and to remove the cover of unit 6. The check for the 10.4- to 10.5-mc output can then be accomplished.

A number of failure reports received by the Bureau of Ships indicates that most frequency errors in the transmitter output can be traced directly to the drift in the frequency output of unit 6. The check described above should therefore be performed at least twice weekly, or as often as required through your experience with the equipment.

6. Check the 1.6- to 2.5-mc output of the phaselocked oscillator of unit 8. This check may be made through the use of the oscilloscope connected to the 100 kc-step jack on the RFO front panel. A multimeter may also be used at the test jack, and it will provide a reading of approximately 0.8 volts to indicate proper operation.

7. If only unit 1 gives an output among the signals checked, perform checks on the regulated filament and B voltage supplies at the metering socket, E-2919 (fig. 12-12). The socket is mounted on the left side of the RFO drawer and is accessible without removing the RFO protective cover.

8. After it is determined that unit 6 is providing its proper outputs, check the output of all the other RFO subunits by measuring the voltage at the test points provided on each unit.

Low r-f output at unit 5 or 9 does not necessarily indicate trouble in the particular unit. A decrease in r-f output may be due to the lack of sufficient drive from one of the inputs feeding the units. Therefore, rather than realigning these units for an increase in r-f output, check the inputs fed to the mixer tubes of units 5 and 9 at their respective sockets.

9. Check the band switch on the RFO front panel, and the cabinet interconnecting cable, which carries the r-f output ot the RFA.

The radio frequency oscillator simplified trouble shooting location block diagram (fig. 12-13) shows a method of localizing trouble in RFO. This diagram indicates minimum r-f output measurements at the test point to indicate which unit is at fault. The heavy blocks indicate the unit (or units) and the associated test point. The lighter blocks identify the units in which the trouble may exist.

For example, unit 2 receives an output from unit 1 and applies one of its outputs to unit 4. If 10.0 volts is not measured at test point 1a, unit 1 (in the left block) is at fault. Output 2a of unit 2 is fed to unit 4. If there is no output from unit 4, the block to the left indicates that unit 2 is at fault. Finally, if a voltage of less than 0.88 volts is measured at test point 4a, following the block to the left will show that again unit 2 is probably at fault. When the trouble has been localized to a unit, refer to the instruction book for maintenance information pertaining to that particular unit.

Radio Frequency Amplifier

When a fault is localized to the RFA, visual indications will aid in locating the faulty circuit. These indications will vary with the mode of transmission and the power level of the output. The voltmeter and antenna current meters on the RFA front panel (fig. 12-11) may be used to check the r-f input and output, respectively. The meter selector switch should be in the RF IN (5V) position for checking the input to the RFA. Also, the INT OSC EXT switch on the RFA front panel should be switched to the INT position and the excitation control turned sufficiently clockwise to allow the 5-volt input to the RFA.

The antenna tuning equipment may be eliminated as a source of trouble by connecting the dummy load, DA-91/U, to the RFA output. In this condition, the transmitter bay is operated without radiating any r-f carrier.

A faulty keying circuit in the low level radio modulator (see K6, figure 7-2, Ch. 7) may cause the RFA to remain unkeyed even though the test

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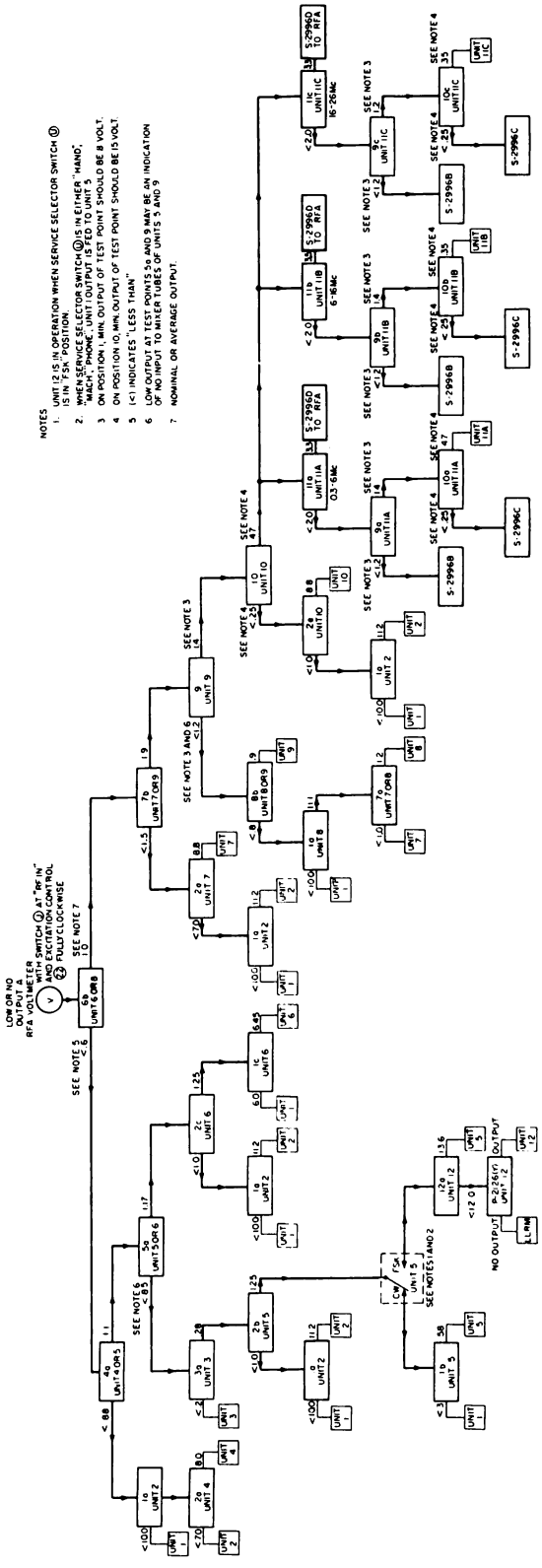
NOTES
1. UNIT 12 IS IN OPERATION WHEN SERVICE SELECTOR SWITCH ⑨ IS IN "FSK" POSITION.
2. WHEN SERVICE SELECTOR SWITCH ⑨ IS IN EITHER "HAND", "MACH", "PHONE", UNIT 0 OUTPUT IS FED TO UNIT 5
3. ON POSITION 1, MIN. OUTPUT OF TEST POINT SHOULD BE 8 VOLT.
4. ON POSITION 10, MIN. OUTPUT OF TEST POINT SHOULD BE 15 VOLT.
5. (<) INDICATES "LESS THAN"
6. LOW OUTPUT AT TEST POINTS 5a AND 9 MAY BE AN INDICATION OF NO INPUT TO MIXER TUBES OF UNITS 5 AND 9
7. NOMINAL OR AVERAGE OUTPUT.
LOW OR NO OUTPUT A RF VOLTMETER WITH SWITCH ⑨ AT "RF IN" AND EXCITATION CONTROL ⑫ FULLY CLOCKWISE
SEE NOTE 7
SEE NOTE 5
SEE NOTE 6
SEE NOTE 3
SEE NOTE 4
SEE NOTES 1 AND 2
NO OUTPUT
OUTPUT
UNIT 1
UNIT 2
UNIT 3
UNIT 4
UNIT 5
UNIT 6
UNIT 7
UNIT 8
UNIT 9
UNIT 10
UNIT 11A
UNIT 11B
UNIT 11C
UNIT 12
S-29960 TO RF-8
S-29996C
S-29968

Figure 12-13. -Radio frequency oscillator troubleshooting block diagram.

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key is held closed. Considerable time may be saved by checking the keying voltage from the LLRM. Assume that the test key (T in figure 12-14) is operated and the transmitter is in standby condition during phone. A multimeter, ME-25A/U (or equivalent) connected between terminal 44 of terminal board E-1397 (chap. 6, fig. 6-2) and chassis ground should read zero volts. With the test key in its center position, the meter should read at least -18 volts, and as much as -30 volts.

If the LLRM is properly producing this voltage, the intermediate power amplifier and the power amplifier stages should be checked by using the IPA current and PA current meters (fig. 12-11) in conjunction with the metering selector switches. Table 12-1 shows readings for all modes of transmission and power levels, with the transmitter set at a frequency of 6 mc and the test key in the operated condition. The antenna tuning equipment is adjusted for the lowest possible standing wave ratio.

The RFA uses voltages that may be dangerous. Extreme care and caution must be exercised when working with or performing maintenance on the RFA.

If the power supply operating voltages (as indicated by the voltmeter readings) are incorrect, check the corresponding power supply. If the IPA and PA operating voltages are correct but there is no IPA or PA cathode current, I_k (500 ma) and I_k (1A) respectively, check the buffer amplifier tube and the LLRM keying circuits discussed earlier. If the cathode current is absent or low in only one of these stages, make a check of that stage.

It has been observed that the blower motor in the RFA occasionally draws an excessive current sufficient to blow the 3-amp fuse on the RFA front panel. When this fuse is opened, the filament voltage for the buffer and intermediate power amplifier is removed (see figure 6-2, chapter 6).

In performing preventive maintenance, clean the filters associated with this blower to avoid overheating and overload of the motor. In any case, try to determine the cause of trouble before the fuse is replaced.

RFA overload frequently occurs in the circuits of the IPA and PA stages. When overload is present, the PA SCREEN or PA PLATE overload indicator lights will go off. Check the power amplifier using the PA current meter in conjunction with the PA switch. An r-f overload indication may be caused by an absence of power amplifier plate voltage, or excessive plate or screen dissipation. Excessive dissipation is usually due to the improper setting of the RFA input excitation control. The reading on the voltmeter (with the voltmeter switch in the RF IN position) should not exceed 5 volts.

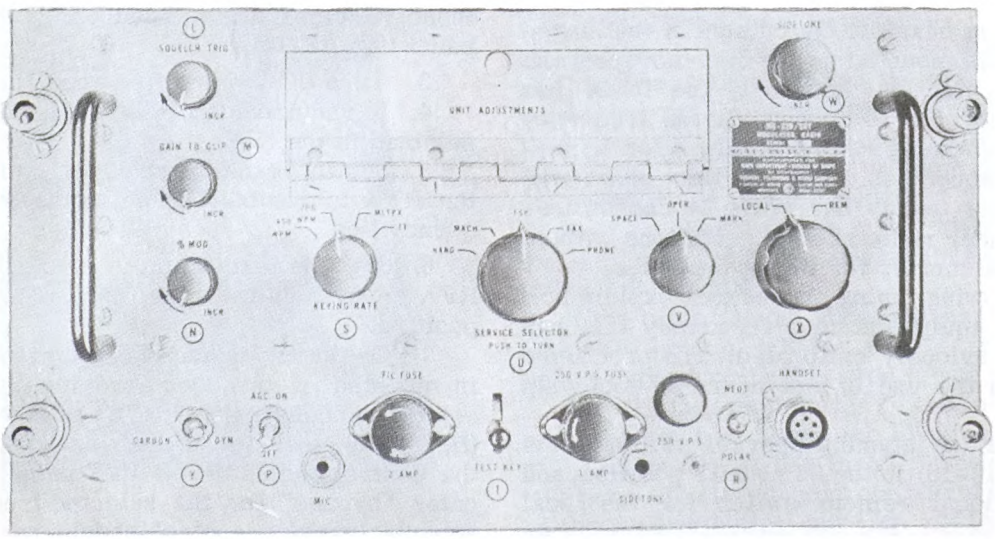

Figure 12-14. —Low level radio modulator front panel controls and indicators.

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Table 12-1.—RFA Test Meter Readings.

Meter		Meter selector switch position	100-watt operation Phone	CW FSK FAX	500-watt operation Phone	CW FSK FAX
Volt meter	Volt meter	RF IN (5 V)	2*	2*	2*	2*
		BIAS (500 V)	-200	-200	-200	-200
		LV (500 V)	260	260	280	280
		MV (1000 V)	430	430	460	470
		PA Ec2 (1000 V)	220	220	390	400
		PA Eb (5000 V)	900	1100	2300	2800
IPA current	IPA (G)	Ic1 (5 ma)	1	1	1	1
		Ic2 (10 ma)	7.5	7.5	9	9
		I_k (500 ma)	60	60	60	60
PA current	PA (H)	Ic1 (50 ma)	18	20	15	12
		Ic2 (100 ma)	68	54	28	16
		I_k (1A)	0.2	0.2	0.3	0.3

* Depends upon setting of excitation control (ZZ)

Tuning the Transmitter

The tuning chart on the outside of the hinged flap, which is mounted on the door covering the frequency selection knobs on the RFO, has tabular space for recording all the frequency selection and tuning control settings for 10 preselected frequencies. Use of this chart will permit tuning to any one of the 10 chosen frequencies under radio silence conditions without energizing a carrier for tuning purposes.

The following tuning procedures explain how to tune the equipment to a frequency that has not been previously calibrated. The procedure is based on the use of a standard 35-foot whip antenna.

1. Place the standby operate switch on the LVPS (fig. 12-16) in the STANDBY position, and check the local remote switch for the local position.

2. Select the desired mode of transmission: either hand-key, machine-key, FSK, facsimile, or phone. The controls for selecting any one of these services are located on the LLRM front panel (fig. 12-14).

3. Turn the transmitter group power on.

4. If required, the new frequency should be selected in the RFO.

5. Set the bandswitch (6 in figure 12-11) in the RFA to the band in which the selected frequency lies.

6. Set the excitation control (ZZ) on the RFA front panel to the maximum clockwise position.

7. On the underside of the hinged flap, which is mounted on the door covering the frequency selection knobs on the RFO, is a tuning chart (fig. 12-9) used for presetting the controls on the control indicator. If the tuning chart indicates "bypass" for the selected frequency, the antenna transfer control should be set to "bypass," and no further tuning adjustments will be required on the control indicator. For other

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frequencies, set the transformer control (AC in figure 12-9) and the antenna coupler loading control (AD) to the positions indicated under the XFMR and COUP columns, respectively, for the frequency being transmitted. The antenna transfer control (AD) should be set to the TUNER-IN position, and the SWR calibrate control to its 8:1 position.

8. Set both the input and output taps on the load adjusting unit (fig. 12-15) to position 4.

9. Place the standby-operate switch (PP on the LVPS in figure 12-16) in the OPERATE position.

10. Operate the test key (T) on the low level radio modulator (fig. 12-14) to its locking position, which will energize a carrier.

11. With the voltmeter selector (J) in figure 12-11 in the RF IN position, the voltmeter reading should not exceed 5 volts. If it does, adjust the excitation control (ZZ) to reduce the reading to 5 volts maximum.

12. Loosen the dial lock of the tune IPA control, and operate the control for maximum output of the buffer and IPA stages. Optimum tuning of these stages is indicated by maximum PA grid drive, as shown by a maximum reading on the PA current meter in the RFA with the PA meter selector switch (H) in the Ic1 position. In tuning, note the two peaks on the PA current meter. The tune IPA control (B) should be set to the peak at the lowest tune IPA dial reading to insure that the tuning is to the desired frequency rather than the second harmonic. Tighten the dial lock.

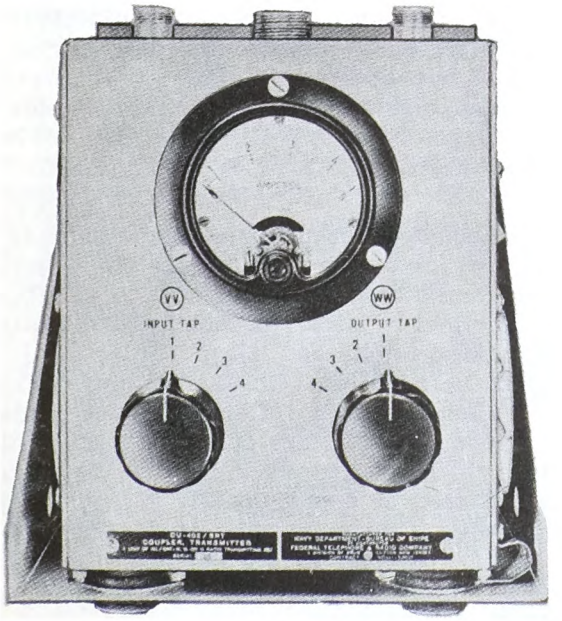

Figure 12-15. —Load adjusting unit, front panel controls and indicators.

13. Loosen the dial lock of the tune PA control (D) and adjust the control for the maximum output of the PA stage. Maximum tuning of this stage is indicated by maximum reading on the PA current meter with the PA meter selector switch (A) in the Ic2 position. As with the tune IPA, the tune PA control (D) should be set to the peak occurring at the lowest tune PA dial reading. If the meter reading goes off scale, adjust the excitation control (ZZ) to bring the meter reading back on scale.

14. On the control indicator (fig. 12-9) depress the up button (AF) until the position meter reads 100. Depress the down button (AG) until the indicator on the SWR balance meter shows a dip toward the green area. GENERALLY, THE POSITION METER SHOULD READ ABOVE 50 FOR ALL TUNNING PROCEDURES, AND ESPECIALLY FOR THOSE FREQUENCIES FROM 0.5 MEGACYCLES, and HIGHER.

The calibrations, 8:1, 4:1, and 2:1, on the SWR calibrate control denote that (with a reading of zero at the center of the scale on the SWR balance meter) the standing wave ratio on the line will be either 8:1, 4:1, or 2:1, depending on the setting of the SWR calibrate control. A reading in the green area of the meter indicates a SWR lower then the chosen position of the SWR calibrate control, and a reading in the red indicates a standing wave ratio higher than the selected position of the control.

Alternately operate the up (AF) and down (AG) buttons, while holding the slow button (AH) depressed until a maximum left-hand deflection is found. As this is done, the SWR calibrate control (AB) should be moved from the 8:1 to the 4:1 position for a more sensitive tuning indication. If the SWR balance meter still reads near the left-hand end of the scale, change the setting to the 2:1 position.

If the tuning chart designates that the antenna transfer control (AE) be set to BYPASS, the tuning procedure in this step is not required.

15. If, upon completion of the tuning adjustments described above the SWR balance meter still indicates a higher than 2:1 ratio, adjust the positions of both the input and output tap switches on the load adjusting unit (fig. 12-15) to improve the reading. If no improvement is seen, return both controls to the initial position 4.

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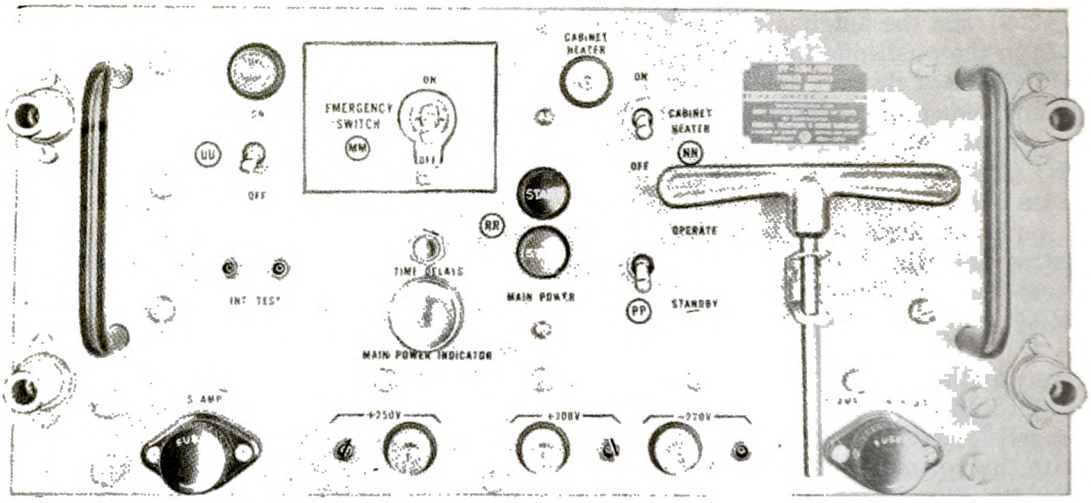

Figure 12-16.—Low voltage power supply, front panel controls and indicators.

16. The tune PA control (D) on the RFA (fig. 12-11) should be "touched up" for optimum reading on the r-f ammeter on the load adjusting unit (fig. 12-15). This action compensates for the changes in the output impedance caused by the tuning of subsequent components after the first setting of the tune PA control. Tighten the dial lock after re-adjustment is accomplished.

17. With the PA meter selector (H) in the Ic2 position, check the PA screen grid current. If the reading is in excess of 70 ma, adjust the excitation control to reduce the value to 70 ma. If the reading is below 70 ma, and if the excitation control is not fully clockwise, increase the control to approach a reading of 70 ma, making sure that the RF IN reading on the voltmeter does not exceed 5 volts. If operating at the 500-watt level in an AN/SRT-15 or AN/SRT-16, the PA Ic2 reading should be checked again for possible necessity of making further readjustments of the excitation control.

18. Restore the test key (T) on the LLRM (fig. 12-14) to its normal off position. If the equipment is to be operated either in HAND or PHONE and the automatic bypass feature is desired, place the antenna transfer control (AE) in figure 12-9 in the remote position after tuning is completed. This allows automatic bypassing of the antenna tuning equipment during periods of no transmission for purposes of using a receiver with the same antenna being used for transmission.

Occasionally, while you are performing step 14, standing waves of sufficient magnitude may be developed between the antenna coupler, r-f tuner, and antenna to break down the insulation of the connecting cable. This causes a corona effect from the cable to the surrounding air, which punctures the coaxial line. Further tuning of the transmitter cannot be accomplished (as noted on the SWR balance meter) until repair of the defective line is completed.

To determine the point of corona on the line, an observer should be placed on the weather deck while the transmitter is being keyed. Corona is evidenced by an electric discharge, which punctures the polystyrene insulation at a point of high voltage. AFTER THE POINT OF DEFECT IS DETERMINED, THE TRANSMITTER SHOULD BE TURNED OFF AND PROPERLY TAGGED.

The technician should obtain permission from the officer of the deck to secure all power to the particular mast involved and to climb the mast to effect repairs. He should also take along a moistened cloth and clean the area of the cable to be repaired. He should also carry a roll of electrical tape. After the area is cleaned, one or two wraps of tape should be sufficient to restore the insulation properties. Place the tape over an area of approximately 2 inches from either side of the point of puncture. The use of glyptal or petrolatum spread over the entire cable serves as an additional protection to prevent future breakdown.

The transmitter should now be turned on and keyed to determine if the repair procedure was adequate. If repair cannot be accomplished in

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this way, replacement of the length of defective cable may be necessary.

Low Level Radio Modulator

When a trouble has been localized to the low level radio modulator, first determine in which major circuit the fault exists. To simplify trouble location, the LLRM is divided into three major circuits: audio circuits, keying circuits, and power supply circuits.

When the trouble occurs in the audio circuits of the LLRM, the transmitter will not modulate properly during phone operation. It should be remembered that when operating at the 500-watt level, the keying signals are passed through the LLRM to the HLRM. Therefore, if there is no modulation present during either 100- or 500-watt operation, the oscilloscope (such as the OS-8A/U or AN/USM-32) should be used to locate the faulty unit of the audio circuits.

Because of the high current passing through the cathodes of the push-pull audio power amplifiers during phone operation, the cathode resistors have been found frequently to burn out (see figure 7-2, chapter 7). To minimize the recurrence of this trouble, it is suggested that the service selector switch (on the front panel of the LLRM) should not be placed in the PHONE position unnecessarily.

When the trouble occurs in the keying circuits of the LLRM, the transmitter will not operate properly in the HAND, MACHINE, and FSK positions of the service selector. The contacts of all relays in the AN/SRT-14, 15, and 16 equipments occasionally become dirty, causing sluggish operation. This is particularly true of the keying and antenna switching relays. The operation of these relays is discussed in chapter 7 of this training course.

A time delay is incorporated in the transmitters bearing serial numbers above 931, which has considerably reduced failure of the keying and antenna switching relays. However, because of the heavy work load imposed on the equipment, all relays should be frequently cleaned with a burnishing tool to reduce the probability of arcing at the contacts. Burnishing the contacts decreases the contact resistance.

The infrequent recurrence of failure reports concerning the high level radio modulator indicates that very little transmission time is lost as a result of trouble in this unit. The procedure outlined in the POMSEE program for the HLRM should be sufficient to insure proper operation.

Power Supplies

Fuses blown, because of the momentary overloads and fluctuating supply voltage, account for the greatest percentage of all failures in the power supplies. When a fuse is to be replaced, the equipment should be turned off. To insure proper operation, follow the maintenance procedure as outlined in the POMSEE program.

ALL POWER SUPPLIES USE VOLTAGES WHICH ARE DANGEROUS. EXTREME CARE SHOULD BE EXERCISED WHEN SERVICING ANY OF THE POWER SUPPLIES.

COMMON RECEIVER MAINTENANCE PROBLEMS

PREVENTIVE MAINTENANCE

Radio receivers AN/SRR-11, 12, and 13 are composed of 12 assemblies: antenna, r-f mixer, oscillator, first i-f, second i-f, audio, BFO, crystal calibrator, power supply tuning dial, and filter. Most of these assemblies can be removed and replaced by loosening a few captive screws. The major assemblies are divided into subassemblies, many of which plug into the assembly. Each plug-in assembly contains a subminiature electron tube and its associated parts.

Plug-in subassemblies are of two types: plug-in boards and plug-in units. The plug-in boards are utilized in the antenna, r-f mixer, and oscillator assemblies. The plug-in units are employed in the first i-f, second i-f, audio, BFO, and crystal calibrator assemblies. The power supply assembly uses conventional type tubes, and does not contain subassemblies.

Changing Tubes

Performing proper preventive maintenance procedures in accord with the POMSEE manuals (supplied with the equipment) will avoid needless breakdown of the receivers. Failure reports show that the major cause of breakdown in the AN/SRR-11, 12, and 13 receivers is defective tubes. This type failure should be expected. However, checking tubes in these equipments as a periodic maintenance is not recommended. When failure occurs, the proper troubleshooting procedures (as described later) will determine

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which tube may have failed. As long as the performance of the equipment is satisfactory, the subminiature tubes should not be disturbed.

Spare assemblies should be kept and stored in a locker that is free of excess moisture. Never stock unpacked units on top of each other.

Making A Sensitivity Check

The operation of the AN/SRR-11, 12, and 13 receivers is considered in chapter 4 of this training course. It is assumed that you have read that chapter and are familiar with its contents.

During the preventive maintenance procedure, a check of the receiver sensitivity is recommended. For the A1 BROAD and A1 SHARP Positions of the reception control (fig. 12-17), and FSK position on the AN/SRR-11, the procedures follow.

1. With the reception control set to the A1 BROAD position, disconnect the antenna simulator from the signal generator (AN/URM-25 series) and connect a jumper across the antenna input terminals.

2. Hold the add decibels switch to the -10 position, and adjust the gain control for -10 db on the output meter. Remove the jumper and connect the signal generator to the receiver input through the antenna simulator.

3. Set the signal generator for an unmodulated output, and adjust the generator frequency for maximum indication on the tuning meter.

4. Set the reception control to the A1 SHARP position, and adjust the frequency vernier control for maximum indication on the output meter.

5. Set the reception control to A1 BROAD. With the add decibels switch at 0, adjust the generator output level for 0 db on the output meter.

6. Upon completion of step 5, record the signal generator output level reading as a measure of the receiver sensitivity. Table 12-2 shows the minimum allowable readings that should be obtained for the various frequencies of the respective receivers with the reception control in the A1 BROAD position.

A series of poor sensitivity checks may sometimes indicate poor receiver alignment. However, failure of some components of the receiver gives the same indication as misalignment. The alignment procedure should be attempted only when the possibility of component failure has been eliminated, and the need for alignment has been established.

Table 12-2.—Receiver Sensitivity Reception Control Set to A1 BROAD Position.

Receiver type	Frequency range	Minimum reading
AN/SRR-11	14 to 18 kc	8 μv
	18 to 100 kc	5 μv
	100 to 600 kc	3.5 μv
AN/SRR-12	0.250 to 8.0 mc	5 μv
AN/SRR-13	2 to 16 mc	6 μv
	16 to 32 mc	10 μv

CORRECTIVE MAINTENANCE

Localizing the Trouble

When trouble occurs, or when it is desired to make a systematic check of the receiver performance, visual inspection of the chassis is often helpful. Check for charred insulation, discoloraton of parts, leakage of potting compound, and for other indications of abnormal operation. If the parts appear normal and odor of burned insulation is not detected, a further check to localize the source of the trouble will be necessary.

The manner in which the receiver functions (or fails to function) should be helpful in localizing the trouble to a particular assembly or subassembly. Observe the lamps located on the front panel of the receiver. If the dial light is glowing after the power has been turned on, the power input and power transformer primary circuits are functioning properly. If the dial lamp does not glow, switch the lamps control to the SPARE position. Failure of the lamp to light after this procedure may indicate a blown fuse.

If the pilot light is glowing, the d-c power supply is operating. If the light does not glow, check the tubes in the power supply.

Stage Gain Measurements

By performing stage gain measurements, low receiver output can be localized to a particular stage. Since the subminiature tubes are

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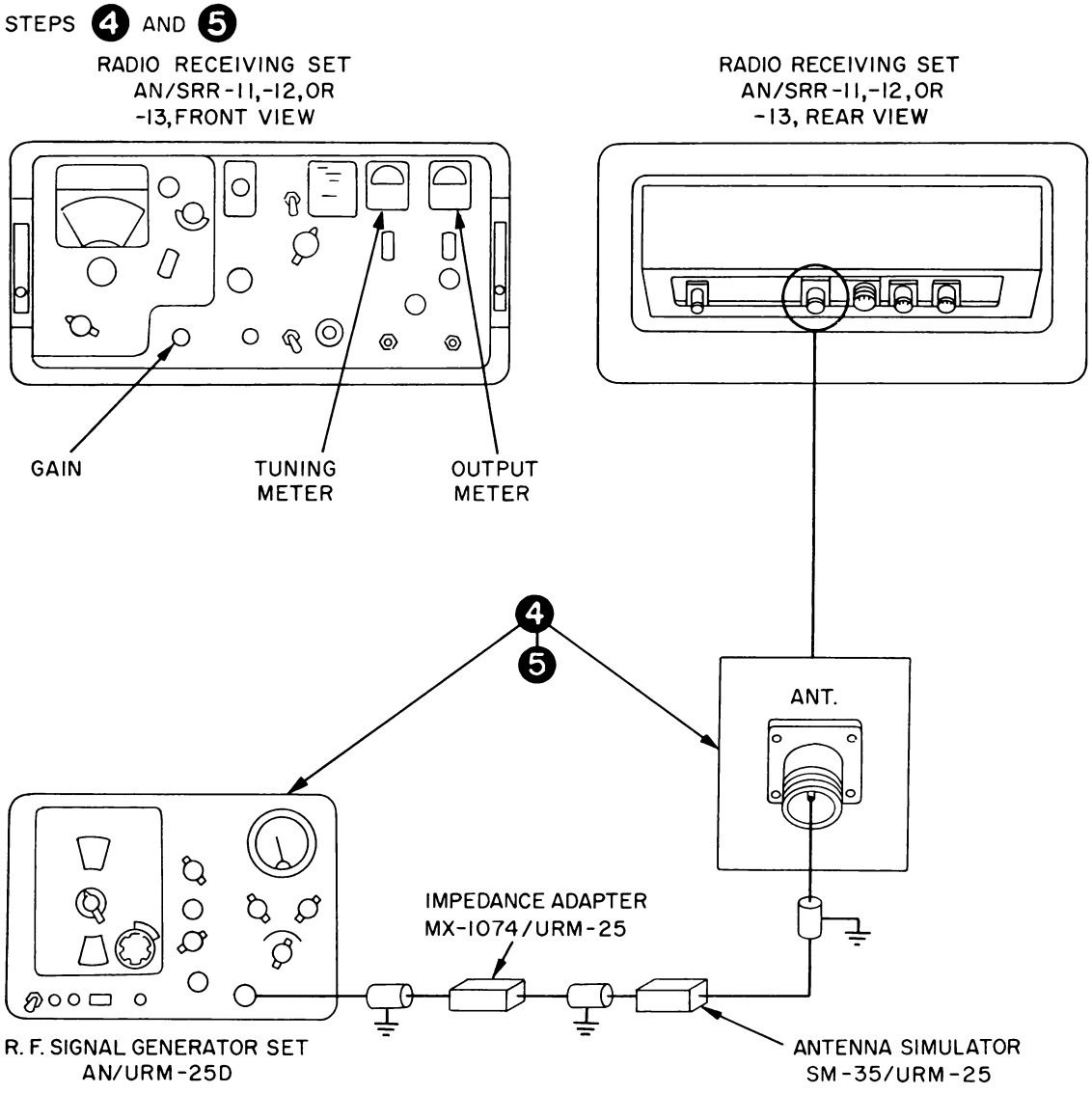

Figure 12-17. —Test setup for making sensitivity check.

soldered into the subassemblies, and cannot be conveniently tested in the usual manner, stage gain measurements may be used in locating faulty tubes.

The test setup for making stage gain measurements of the audio system is shown in figure 12-18. The check procedure and test setup must be altered in accordance with the circuit requirements (see section 7 of *Radio Receivers AN/SRR-11, 12, 13,* NavShips 91875A). The output of the r-f signal generator or audio generator is applied to the stage under test. The 1.0-μf capacitor is used, as stated in note "C" (fig. 12-18).

All controls should be set in the appropriate position for making the required check. When making stage gain measurements of the r-f and i-f stages, particular care must be taken to place the band selector and tuning dial of the receiver, and the modulation controls of the signal generator to their correct settings. A good knowledge of the receiver circuitry will be helpful in determining these settings.

Stage gain measurements accomplished on the audio assembly should yield the readings shown in table 12-3. In all cases, the signal generating source should be applied to the input of the stage. Output signals at the output of the

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receiver may be read either from the output meter or from a voltmeter (ME-25/U series).

If a measured stage gain does not agree to within approximately 30% of the value given in table 12-3, the stage should be replaced. If this action fails to correct the trouble, take voltage and resistance readings of the stage to determine if the d-c potentials are low or if an open circuit exists.

Checking Tubes Mounted in Subassemblies

Test Adapter MX-2012/U (fig. 12-19) is a wiring adapter to enable testing of subminiature tubes mounted on the subassemblies, without removing the tubes. The unit is designed to be used with tube testers of the TV-3/U series (fig. 12-20) for testing tubes in several Navy radio receivers, including types AN/SRR-11, 12, 13, and 13A. A schematic diagram of the MX-2012/U is shown in figure 12-21.

The sockets mounted on the front panel of the test adapter provide the connections to the various pins on the subassemblies containing the tubes. The Octal plug (fig. 12-19) of the adapter is connected to the octal socket of the tube tester. Operating potentials are applied through the cable from the tube tester to the test adapter via P101 (fig. 12-21).

Sockets E101 and E102 provide direct connections for tube contacts of the i-f and a-f plug-in units. J101 and J102 are the receptacles for the r-f plug-in units.

The individual socket assemblies are wired so as to connect the elements of the tubes, in the subassemblies being tested, to the corresponding circuits of the tube tester. Thus, the tubes can be tested using the normal tube tester functions. The four jumper cables (fig. 12-19) complete the circuits in cases where components on the subassemblies interfere with the normal connecting procedures.

Pushbutton switch S101 (fig. 12-21) is depressed to complete the biasing circuit between pins 5 and 6 of the octal socket. The button is depressed to complete these circuits at the same time the tube tester button is depressed when testing some assemblies. Consult the tube tester, TV-3/U instruction book, NavShips 92743 for circuit operation and the necessary tube test instructions.

The two power supply diodes (type 6x4) may be checked without the use of the MX2012. Both of the parallel rectifiers should be replaced at the same time. Replacement of the defective tube only will cause a serious mismatch in the amount of load current carried by the separate tubes. Through this action, the useful life of the new tube is shortened.

Binding posts E103 and E104 are provided for testing of type 5644 voltage regulator tubes. These tubes are unmounted, and tested by connecting wire leads to the binding post on the panel, after setting the tube tester to the values stated on the tube tester chart. Pushbutton P5 on the tube tester front panel (not shown) is depressed when making this test. When P5 is depressed, +150 volts exist between the binding posts E103 and E104. If the needle on the tube tester deflects in the wrong direction, reverse the leads at the binding posts.

Checking the Crystal Calibrator Frequency

The crystals employed in the crystal calibrator circuits are operable in the temperature range from -40 to +85 degrees Centigrade. The circuits of the crystal calibrator of the AN/SRR-11, 12, and 13 receivers are discussed in chapter 4 of this training course.

To ascertain that the crystal calibrator circuits are on frequency, a very accurate frequency meter source (such as provided by the LR series) is required. The reception control should be placed in the A1 position to allow the operation of the beat frequency oscillator. Also, remove the cover from the crystal calibrator sub assembly.

The frequency meter should be tuned to the frequency of the crystal calibrator output, as shown in table 12-4. Refer to chapter 4 for the components listed and the point of connection of the frequency meter. If the frequency meter used cannot be tuned to the fundamental frequency, obtain the nearest harmonic of the crystal calibrator frequency.

The frequency meter output is applied at the point of the crystal calibrator output to the i-f amplifier. The two signals reinforce each other, and are heterodyned with the output of the beat frequency oscillator. Both signals should yield a zero beat at the same setting of the BFO frequency vernier control. If the zero beats do not occur simultaneously, adjust the crystal calibrator tuning capacitor until an absolute zero beat occurs.

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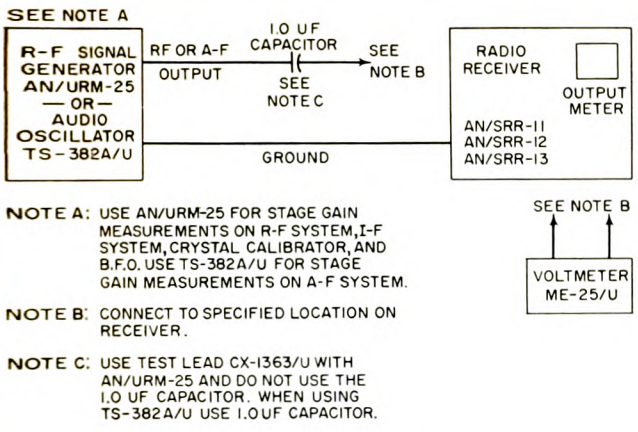
SEE NOTE A
R-F SIGNAL GENERATOR AN/URM-25 —OR— AUDIO OSCILLATOR TS-382A/U
RF OR A-F OUTPUT
1.0 UF CAPACITOR
SEE NOTE C
SEE NOTE B
RADIO RECEIVER
OUTPUT METER
AN/SRR-11
AN/SRR-12
AN/SRR-13
GROUND
SEE NOTE B
VOLTMETER ME-25/U
NOTE A: USE AN/URM-25 FOR STAGE GAIN MEASUREMENTS ON R-F SYSTEM, I-F SYSTEM, CRYSTAL CALIBRATOR, AND B.F.O. USE TS-382A/U FOR STAGE GAIN MEASUREMENTS ON A-F SYSTEM.
NOTE B: CONNECT TO SPECIFIED LOCATION ON RECEIVER.
NOTE C: USE TEST LEAD CX-1363/U WITH AN/URM-25 AND DO NOT USE THE 1.0 UF CAPACITOR. WHEN USING TS-382A/U USE 1.0 UF CAPACITOR.

Figure 12-18. —Test setup for stage gain measurements.

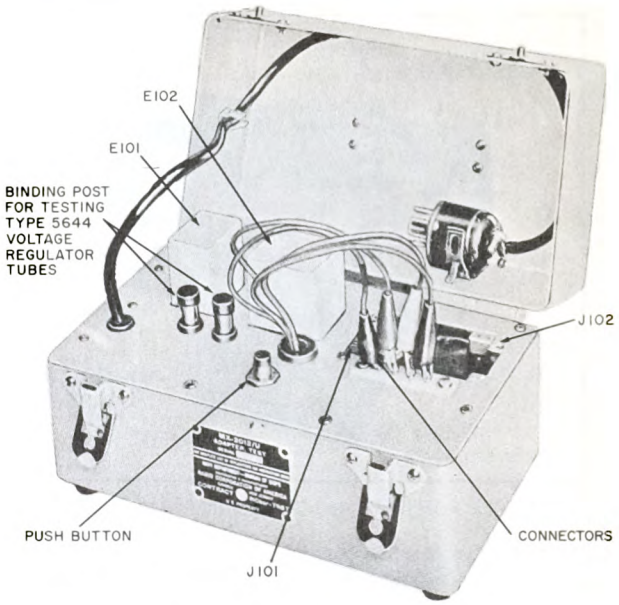

Figure 12-19. —Test adapter MX-2012/U.

Table 12-3.—Stage Gain Measurements.

		INPUT		OUTPUT	
Unit	Frequency CPS	Location	Level volts	Level decibels	Location
AF	1000	Control grid of output amplifier	1.5	Odb (6 mw)	Output meter
	1000	Control grid of third amplifier	0.6	Odb (6 mw)	Output meter
	1000	Plate of second limiter	0.6	Odb (6 mw)	Output meter
	1000	Cathode of first limiter	0.6	Odb (6 mw)	Output meter
	1000	Control grid of second amplifier	0.05	Odb (6 mw)	Output meter
	1000	Plate of silencer diode	0.05	Odb (6 mw)	Output meter
	1000	Output from first amplifier coupling network	0.055	Odb (6 mw)	Output meter
	1000	Output of first audio amplifier	0.25	Odb (6 mw)	Output meter
	1000	Control grid of first amplifier	0.017	Odb (6 mw)	Output meter

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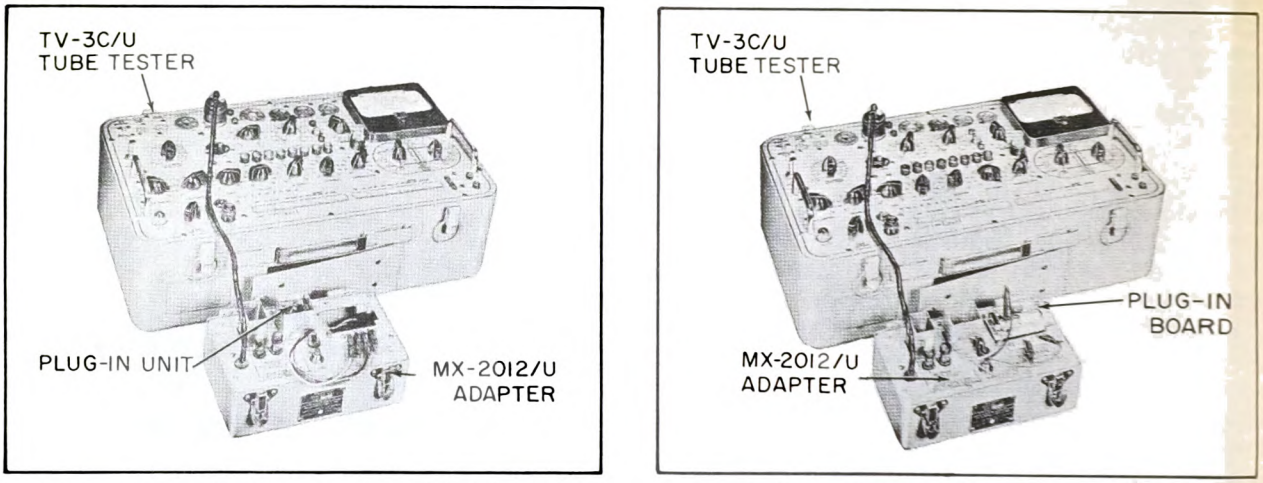

Figure 12-20.—Setup for testing tubes in plug-in units and plug-in boards.

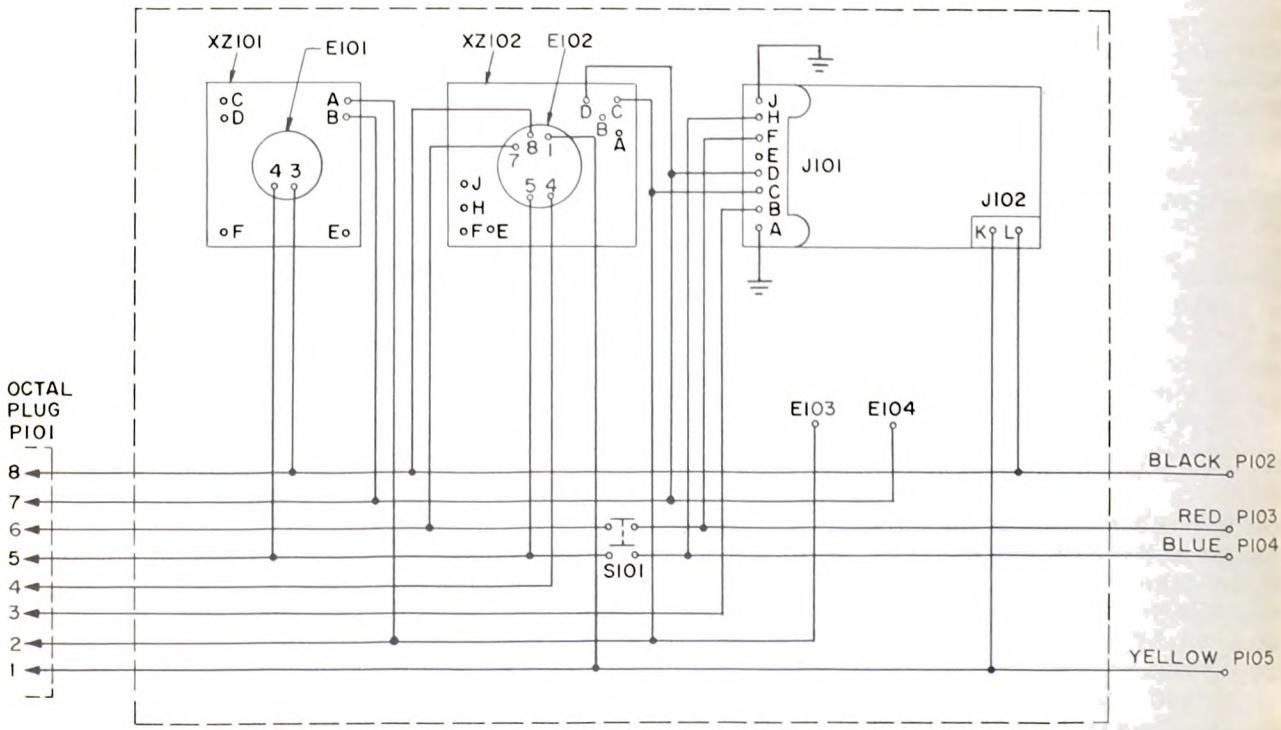

Figure 12-21.—Test Adapter MX-2012/U, Schematic diagram.

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Table 12-4.—Crystal Calibrator Alignment.

Equipment	Crystal calibrator fundamental frequency	Frequency adjustment	Frequency Meter connection
AN/SRR-11	50 kc	C88 figure 4-8	Control grid of 2nd IF ampl figure 4-6
AN/SRR-12	50 kc	C1 figure 4-13	Control grid of 2nd IF ampl figure 4-16
AN/SRR-13	200 kc	Same as AN/SRR-12	Same as AN/SRR-12

Maintenance of Transmitting and Receiving Switchboards

The receiver transfer switchboard, SB-2/SRR (fig. 12-22) and the Transmitter Transfer Switchboard, SB-83/SRT (fig. 12-23) both operate on the principle of cross-mat parallel wiring of their respective switches. The receiver switchboard has five vertical rows of ten double-pole, single-throw switches, continuously rotable in either direction. The transmitter switchboard has five vertical rows, (two shown) of ten twelve-pole, single throw switches, also continuously rotable in either direction.

The wiring plans of both switchboards are similar. One side of each of the receiving switches within a vertical row is wired in parallel with the same sides of the other nine switches within that row. Similarly, the sides of each switch that is not wired in parallel vertically is wired in parallel horizontally with the other sides of each of the other four switches in the same horizontal row.

The receiver switch panel carries the receiver output only, as was true with the now obsolete receiver transfer plug panel. However, the transmitter switch panel (fig. 12-23) operates both start-stop circuits (including keying and indicator circuits) and the transmitter audio control (including microphone voltage, carrier control, and carrier indicator circuits).

All external wiring is brought to the panel through terminal tubes, and is connected to a terminal block located at the back of the cabinet. The numbers on the terminal block from 1 through 6 control the start stop feature of the transmitter, and the numbers 7 through 12 control the audio circuits. The panels themselves are designed as units, and as many units as are required for a particular installation may be mounted together on a common bracket. To facilitate interconnections for group installations, the Bureau provides two types of harnesses for each type panel, one for the vertical, and one for the horizontal interconnections.

To perform wiring or maintenance procedures on either of the switchboards singularly, remove the two screws at the top of the panel. Tilting the panel forward from the top permits access to the wiring connections. Where several panels of the same type are employed, a bank connecting arrangement is used. In such cases, it is necessary to remove the appropriate bolts on the bank in question to tilt the entire bank forward. Care should be exercised to avoid damage to the control knobs when the panel is dropped.

Cold solder joints and dirty switch contacts are the most common troubles occurring in either of the two switchboards. Reports from the field indicate that screws, washers, bits of solder, and other foreign matter are found lodged in the switches. These conditions may cause considerable cross-talk between the various units connected through the panel.

The panel should be thoroughly cleaned prior to installation. Preventive maintenance operations should include a periodic procedure for dusting and cleaning the switch contacts. Clean out the metal parts found in the panels and loose bits of solder on the terminals. Readjust the

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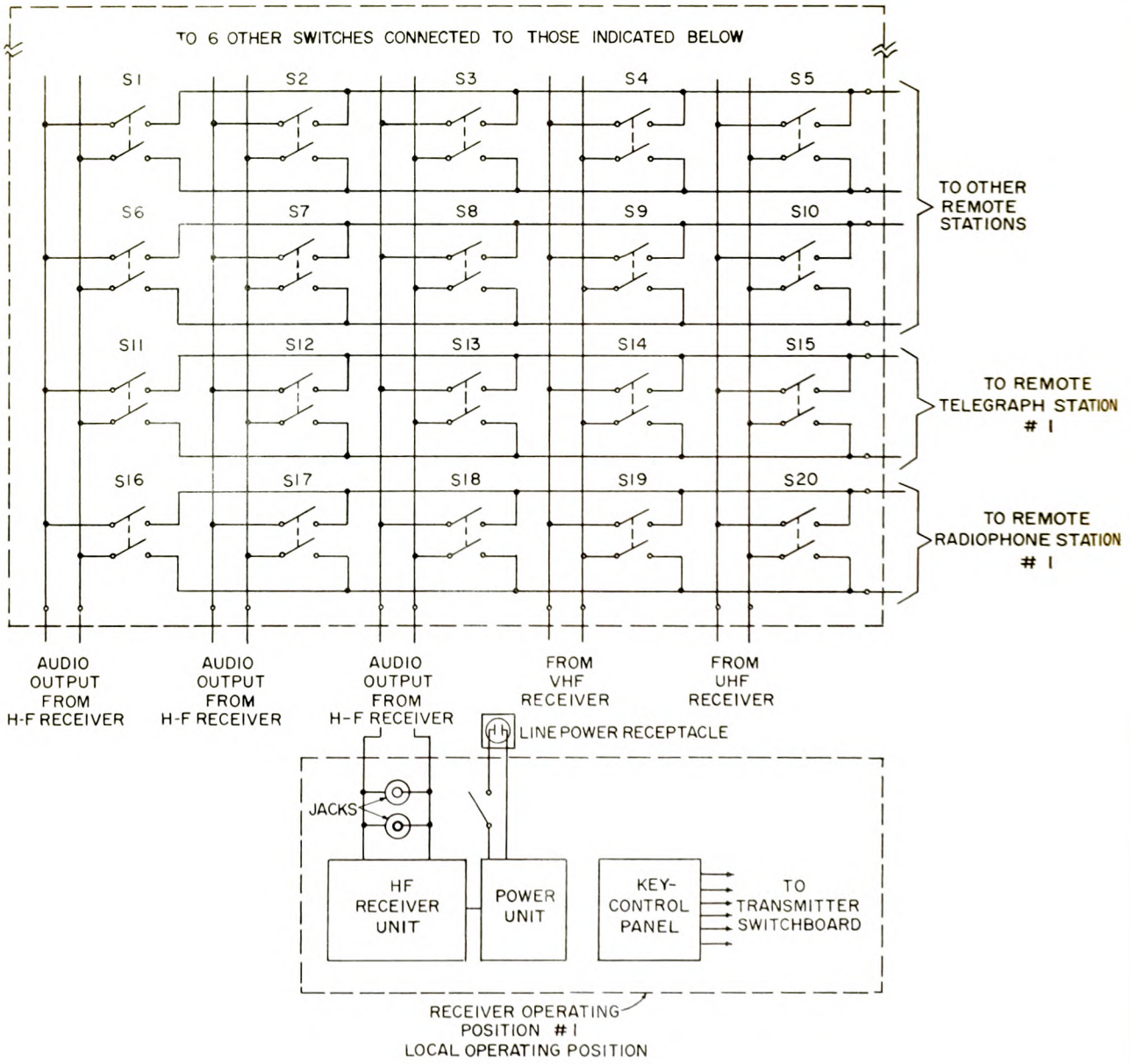

Figure 12-22.—Simplified schematic diagram of the Receiver Transfer Switchboard, SB-82/SRR

knobs on the switches to prevent shorts between adjacent circuits.

Cold solder joints frequently exist in the equipment at the time of installation. These connections are often singled out through visual inspection. Good solder connections are generally bright and firm. Cold solder joints are dull and very easily broken. The ohmmeter is also helpful in locating high resistance connections.

MAINTENANCE OF RADIAC EQUIPMENT

The theory of operation and maintenance of Radiac Set AN/PDR-27G is considered in Chapters 6 and 11 of *Electronics Technician 3*, NavPers 10188-A. This particular radiac equipment

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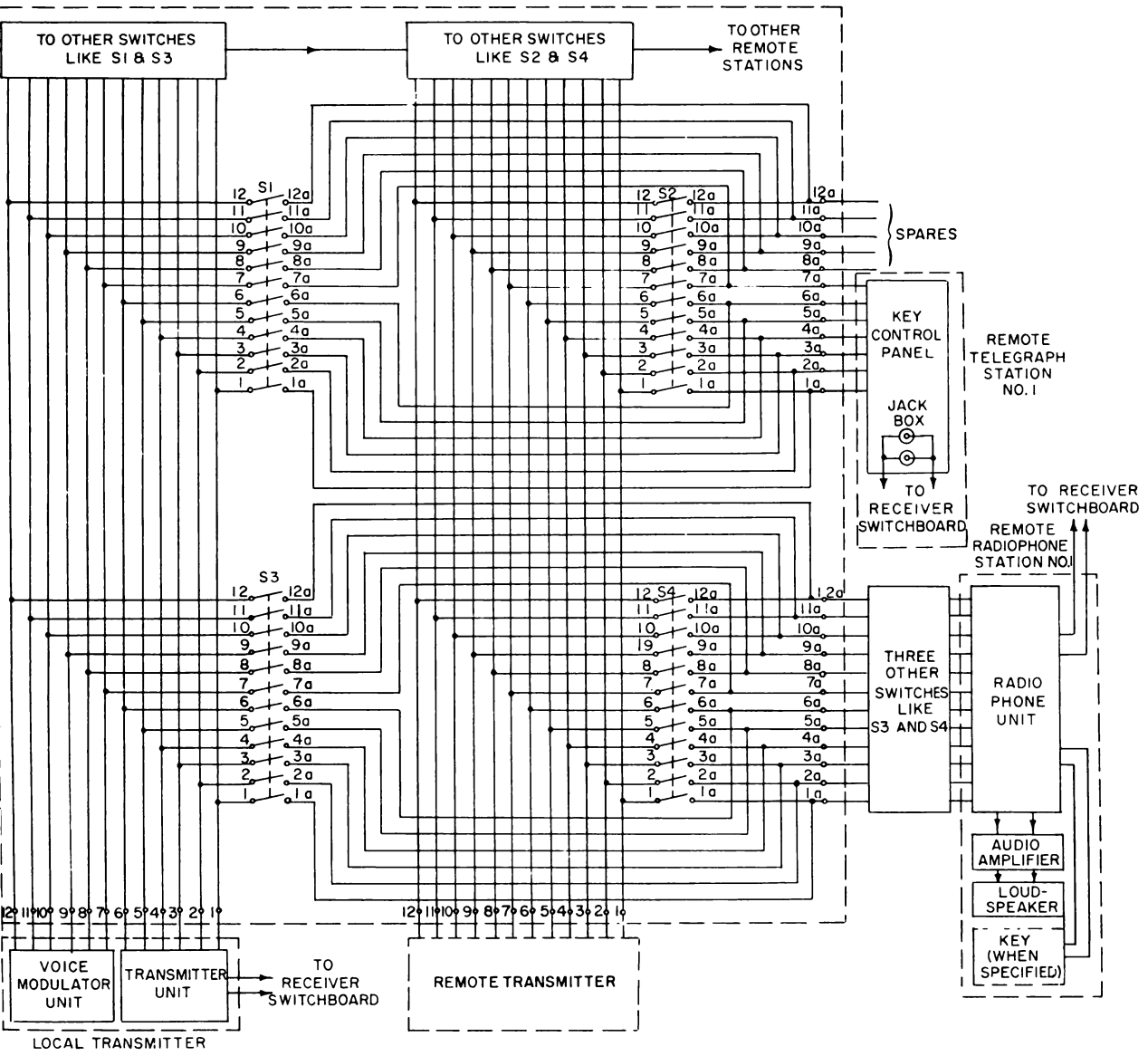

Figure 12-23.—Simplified schematic diagram of the Transmitter Transfer Switchboard, SB-83/SRT.

eceives wide usage in present-day Naval Radiac pplications.

The most com:non maintenance procedures ccomplished on these equipments is the check- ıg and replacement of the batteries and the cor- ection of the radiacmeter sensitivity through ıe recalibration process described in Chapter 1 of the *ET-3* training course.

Since the most common cause of failure is ead batteries, always check the condition of the atteries by turning the meter switch to the BATT COND position when the radiacmeter is inoperative. When the indicating meter pointer rests to the left of the black line marked, BATT, the batteries are depleted and should be replaced. This test, however, only checks the condition of one of the batteries in the equipment. Since it is possible that the other batteries are defective, it is advisable to replace all of the batteries at this time, unless the other batteries have been recently replaced.

The radiacmeter, radiac detector, and headset can be checked with the radioactive test

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sample. This test will yield a qualitative estimate of the performance of the quipment. However, the absolute accuracy of the calibration cannot be determined by this means. The test should be made whenever the existence of trouble is suspected. When it is necessary to perform the full calibration procedure, refer to the *Radiac Set AN/PDR-27G Instruction Book*, NavShips 92071 or to Chapter 11 of the *ET-3* training course.

The calibration of radiac equipment should not be attemped by Electronic Technicians. For the most accurate calibration of the radiacmeter, a larger amount of the radioactive substance than is supplied with the meter is necessary. Thus, accurate calibration of the meter requires the more complete testing and calibrating instruments and the larger radioactive sources, which are available at a radiac repair facility.

QUIZ

1. What program serves as the basis for the recommended preventive maintenance program for all shipboard and shore-based installed electronic equipment?
2. What does POMSEE mean?
3. Where are the equipment Performance Standards Sheets kept so as to be readily accessible at all times?
4. Are the reference column tolerances shown in parentheses in part I absolute limits?
5. Where are the step numbers (enclosed in stars) referenced?
6. How are the test standards compared in order to evaluate the equipments performance?
7. The position meter (fig. 12-9) indicates the position of what antenna tuning element?
8. If the sliding arm of the coil in the r-f tuner is placed too near the ground potential, what will be the effect?
9. If an unused transmitter (AN/SRT-14, 15, or 16) is available, what method of repair can be used to save considerable time?
10. With the test key (fig. 12-14) in the center position (transmitter unkeyed), a multimeter between control grid and ground of the buffer (V1 of figure 6-2) should read approximately what value?
11. If unit 1 of a particular transmitter is known to be operating at the correct frequency, the frequency output of another unit 1 can be ascertained (through the use of Lissajous patterns) in what manner?
12. The frequency output of unit 1 (chapter 5, figure 5-6) can be changed by adjusting what component?
13. When the oscillator tube of unit 3 (chapter 5, figure 5-9) is replaced, what effect generally occurs?
14. The 10-kc output of the blocking oscillator (V4A, chapter 5, figure 5-12) may be observed at what jack on the RFO (fig. 12-9)?
15. Most frequency errors in the transmitter output can be traced directly to the drift in frequency of what unit?
16. The voltmeter and antenna current meter on the RFA front panel (fig. 12-11) may be used for what purpose?
17. The antenna tuning equipment can be eliminated as a source of trouble by connecting what component to the RFA output?
18. The reading on the voltmeter (fig. 12-11) with the voltmeter switch in the RF IN position should not exceed what value?
19. If, when tuning the transmitter, two peaks are noted on the PA current meter while rotating the tune IPA control, B (fig. 12-11) the control should be set to which of the two peaks?
20. Generally, the position meter (fig. 12-9) should read above what value for all tuning procedures?
21. A reading in the green area of the SWR balance meter (fig. 12-9) yields what indication?
22. To simplify trouble location, the LLRM (fig. 12-14) is divided into what three major circuits?
23. Because the subminiature tubes are soldered into the subassemblies and cannot be conveniently tested in the usual manner, what method may be used to locate faulty tubes?
24. Test Adapter MX-2012/U (fig. 12-19) is a wiring adapter used for what purpose?
25. The individual socket assemblies (fig. 12-21) are wired in what manner?
26. Binding posts E103 and E104 (fig. 12-21) are provided for what purpose?
27. The crystals employed in the crystal calibrator circuits are operable within what temperature range?
28. The Receiver Transfer Switchboard, SB-82/SRR (fig. 12-22) and the Transmitter Transfer Switchboard, SB-83/SRT (fig. 12-23) both operate on the principle of what type wiring for their respective switches?

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CHAPTER 13

ELECTRONIC AIDS TO NAVIGATION: LORAN

INTRODUCTION

The loran system is concerned with long range radio navigation. It permits the navigator to determine his position when weather conditions or extreme distance from shore prevent the use of standard methods, such as celestial navigation or radio direction finding.

BASIC PRINCIPLE OF THE LORAN SYSTEM

The AN/UPN-12 Loran Receiving Set, discussed here, is a navigational aid that gives a direct reading in microseconds (at the loran receiver) of the time difference in arrival of signal pulses from master and slave transmitters of a loran transmitting group. In the loran system it is neither necessary (1) to know the exact location of the transmitting station, nor (2) to measure the direction of arrival of the radio signals.

The principle of loran navigation involves establishing a fix at the intersection of two or more "lines of position." (See Chapter 1 of the *Electronic Technician* 3 training course). A loran line of position is a line of constant time difference between the arrival of two pulse modulated r-f signals transmitted from a pair of loran transmitting stations (master and slave transmitters). The stations are usually located several hundred miles apart. Because r-f energy travels at a constant velocity, the time-difference in reception of the two signals is a measure of the difference in the distance from the transmitting stations to the point of observation.

An obtained time-difference reading is used to locate a specific line of position on a special chart. The chart contains various loran lines of position.

If a second time difference is obtained from a second loran transmitting pair, the new time difference is used to locate a second line of position on the chart. The intersection of the two lines of position obtained from the time-difference readings establishes a "fix."

The receiving set of the loran system (fig. 13-1) is basically an instrument for measuring the very small periods of time that elapse between the arrival of signals from the two stations of the loran transmitting pair. To accomplish this, a variable timed interval is generated within the receiving set and displayed on the screen of a cathode-ray tube. When the received signals are positioned at the beginning and end of the timed period, the time difference between arrival of the signals at the location of the receiving set may be read from a revolution-type time-difference counter (fig. 13-1). A second reading obtained in the same manner from a different transmitting pair produces a second time-difference reading on the counter. The first and second-time diffference readings are used to determine geographical location from loran charts and tables (not shown).

IDENTIFICATION OF STATIONS

Because the loran system presents visual intelligence on the indicator screen, the conventional sound method of identifying stations by call letters cannot be used to identify a given pair of loran transmitting stations. Identification is accomplished by using four different transmitting frequencies (known as channels) and by using different pulse recurrence rates. The channel and channel frequencies are channel 1, 1950 kc; channel 2, 1850 kc; channel 3, 1900 kc; and channel 4, 1750 kc. The pulse recurrence rates are divided into three basic groups known as S (special), L (low), and H (high), respectively (see table 13-1).

The pulse recurrence rate for the S group is from 20 to 20 7/25 cps, 25 to 25 7/16 for the L group, and 33 3/9 to 34 1/9 for the H group. Each basic group of pulse recurrence rates

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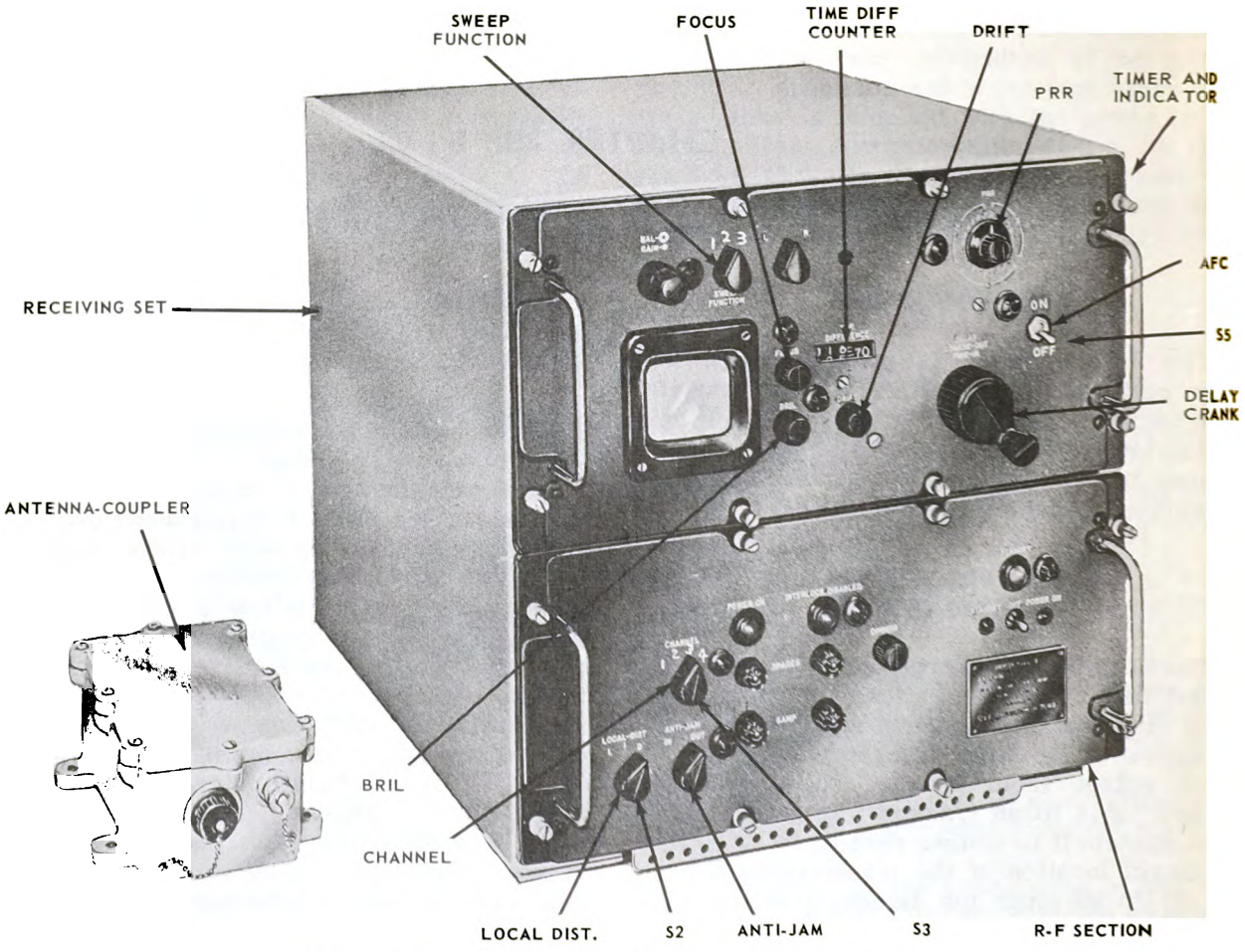

Figure 13-1.-Loran Receiving Set AN/UPN-12.

contains eight station rates designated numerically from 0 through 7. The recurrence interval between pulses (in μs) from the transmitter for the various station rates are also shown in the table.

SELECTION OF STATIONS

To select a given pair of stations it is only necessary to rotate the channel and PRR switches (fig. 13-1) to the appropriate positions. The stations are identified on loran charts according to the r-f channel, the basic pulse recurrence rate, and the station rate. For example, 1H3 designates a station operating on channel 1 with basic pulse-recurrence rate H and station rate 3. An unknown pair of signals may be identified by adjusting the control of the receiver-indicator until the pulses remain stationary, and noting the position of the switches.

LORAN RECEIVING SET AN/UPN-12

BLOCK DIAGRAM

The loran receiving set block diagram, along with some of the major waveforms from various circuits throughout the receiving set, is illustrated in figure 13-2. The general operation of the receiving set is discussed with the aid of this illustration. Frequent reference to the block presentation while studying through this chapter will prove helpful in understanding how the various inputs and outputs of the stages are derived.

The antenna signals are fed into the r-f circuit through an antenna coupler, an antenna tune switch, S1, and an attenuator. The desired

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Table 13-1.—Station Rates and Recurrence Intervals.

Basic Pulse Recurrence Rate Group	Station Recurrence Rate Designation	Frequency (Cycles per Second)	Recurrence Interval Between Pulses (Microseconds)
S	0	20	50,000
S	1	20-1/25	49,900
S	2	20-2/25	49,800
S	3	20-3/25	49,700
S	4	20-4/25	49,600
S	5	20-5/25	49,500
S	6	20-6/25	49,400
S	7	20-7/25	49,300
L	0	25	40,000
L	1	25-1/16	39,900
L	2	25-2/16	39,800
L	3	25-3/16	39,700
L	4	25-4/16	39,600
L	5	25-5/16	39,500
L	6	25-6/16	39,400
L	7	25-7/16	39,300
H	0	33-3/9	30,000
H	1	33-4/9	29.900
H	2	33-5/9	29,800
H	3	33-6/9	29,700
H	4	33-7/9	29,600
H	5	33-8/9	29,500
H	6	34	29,400
H	7	34-1/9	29,300

et of loran stations is selected by means of a hannel control switch, S3 (fig. 13-1), which elects one of four available channels. The nput signals (fig. 13-2) are amplified in the -f amplifier, V1; heterodyned in a crystal ontrolled converter, V2; again amplified in two tages of i-f amplification (V3 and V4) and de- ected in V5A. The input signal level to the -f amplifier stage, V1, is controlled by a hree-position, local distant (L-I-D) switch, S2 fig. 13-1).

The detected pulses from V5A (fig. 13-2) re coupled through video amplifier-limiter '6A and a high pass filter, to video amplifier '6B. A clipper stage, V5B, eliminates negative ortions of the V6A output. Since the V6B output s applied through another video amplifier, V7, o the indicator tube, V60, the action of clipper tage V5B eliminates the portion of the V60 input elow the established reference.

An antijam switch, S4, functions to reduce the effects of jamming by removing low-frequency components of the video signal. High-frequency components are retained to preserve the rise time (leading edge) of the input pulse.

The video output of V6B is coupled along dual paths to the video amplifier, V7, and to video amplifier V10. The V7 output is applied to the indicator tube, V60, as discussed. The V10 output is applied to the A and B gate stages, V11 and V12, respectively, and is effective in controlling the automatic frequency control circuits.

The positive output of the pedestal amplifier, V55, is applied to the A multivibrator, V8, in the frequency control circuits. The frequency control stages synchronize the repetition rate of the receiving set with that of the received signals to hold the pattern displayed on the indicator stationary while readings are being taken. Either automatic or manual frequency control is used, depending upon the setting of the AFC switch, S5.

When AFC is used, the repetition rate of the incoming signals is compared with the receiving set repetition rate in stages V9 through V15. The error voltage derived from the comparison at the discriminator, V15, is used to hold the receiving set pulse rate and the rate of the received signals in synchronism. When manual control is desirable, the AFC switch (fig. 13-1) is placed to OFF. Fine control of the receiving set pulse repetition rate is then determined by the drift control, R86.

The time base of the receiving set is produced by an 80-kc sine wave crystal-controlled oscillator, V17 (fig. 13-2). The oscillator output is shaped in a ringer stage, V18, and clipped in V19 to produce sharp negative driving pulses for a group of 11 counter stages, V20 through V30. Actually 12 counter stages are used. The action of the twelfth counter, V31, is considered presently. The oscillator, V17, is reactance-tube controlled by V16, whose input is the error voltage derived from the comparison of the receiving set repetition rate, and the rate of the received signals at the discriminator, V15, as discussed.

Frequency division is used in the counter stages, V20 through V31, to produce a square wave at each counter output, which is one-half the frequency of its input. Each of the counter stages is an Eccles-Jordan multivibrator.

The square waves at the plates of the third to eleventh counter (V22 to V30) are employed

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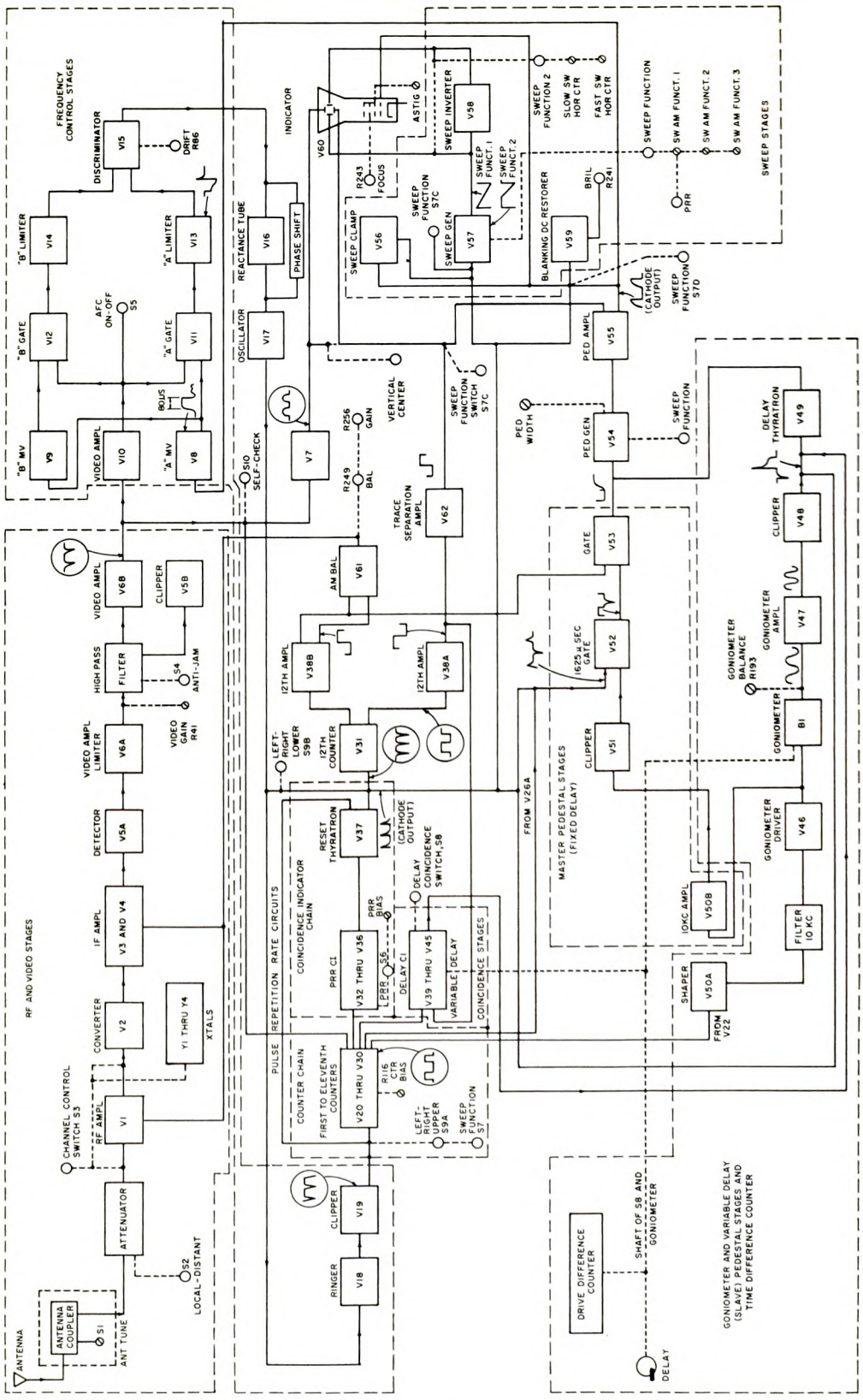

Figure 13-2.—Loran Receiving Set AN/UPN-12, block diagram.

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n the repetition rate and variable delay cir-uits. The 10-kc (100 μs) output of the third ounter is coupled to the shaper, V50A (a part f the variable delay stages), and the 1600-μs utput of the seventh stage, V26, is coupled to ne fixed delay stages.

In the pulse repetition rate circuits the input quare waves from the third to eleventh counter tages (V22 to V30) are combined in a reset nyratron tube, V37, to produce a positive and egative reset pulse when the predetermined ount selected by the PRR switch, S6, is reached. he positive reset pulse is used to trigger all f the counter stages, V20 through V31, thereby esetting the counter stages to start a new ycle. The negative reset pulse is used to drive ne sweep generator, V57, and the blanking d-c estorer, V59.

The recurrence frequency of the reset pulse rom V37 is twice that of the received signals. lowever, each negative reset pulse from the 37 plate (as seen later) triggers the twelfth ounter stage, V31, which produces the final utput frequency of the counter chain. Thus, ince two input pulses are required to produce single output pulse from the counter chain, he period of the output of the twelfth counter, 31, is equal to the period of the received sig-als.

The twelfth (final) counter (V31) plates pro-uce two square waves of opposite polarity, each f which has a period equal to that of the ransmitting group recurrence interval.

The twelfth counter stage (V31) output is lso fed through the twelfth amplifier, V38A, o the trace separation amplifier, V62. The race separation amplifier uses the V38A input o produce a voltage, which positions the elec-ron sweep beam toward the upper section of the ndicator (V60) screen during the period of eception of the pulse from the master trans-nitter of a transmitting group. The total period f this sweep is designated the A period.

During reception of the pulse from the slave ransmitter, the trace separation amplifier V62) output causes the electron beam to be ositioned toward the lower section of the indi-ator tube. The total period of this sweep is alled the B period (see figure 13-3).

The shaper, V50A (fig. 13-2) 10-kc filter nd goniometer circuit comprising V46, B1, V47, nd V48 produces a sine wave from the 10-kc, 00-μs output of the third counter, V22, which s used in the fixed and variable delay circuits. he fixed and variable delay stages, in turn,

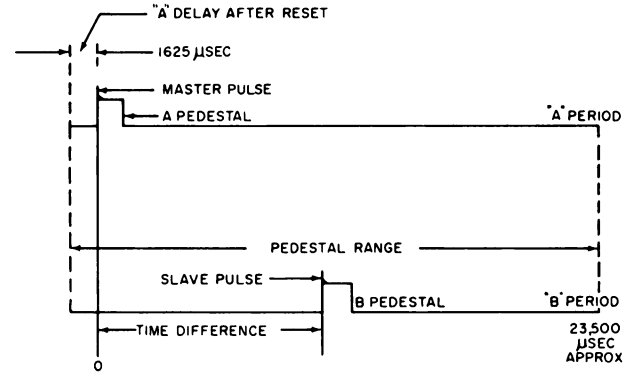

Figure 13-3.—CRT presentation (sweep function 1 operation).

combine their various inputs to produce a trigger pulse, which is coupled to the pedestal generator, V54.

The variable delay voltages from V39 through V45 are combined with the output of the clipper, V48, in the delay thyratron, V49. The fixed delay is derived by combining the output from the seventh counter, V26A, with the output of a clipper, V51, in the 1625-μs gate stage, V52. Clipper V51 is driven by the 10-kc output of V50B.

Two pedestal pulses are produced by the receiving set shown in figure 13-3. One pedestal is produced during the A period, and the other during the B period. The time of occurrance of the A pedestals is fixed by the arrival of the trigger pulse from the seventh counter stage, V26.

A delay crank (fig. 13-1) controls the variable delay system (fig. 13-2) so that the B output pedestal pulse is gated by the input from the twelfth counter, V31, to occur only during the B period. The B pedestal is variable from a time shortly after the reset pulse initiating the the B period to the end of the available delay time incorporated in the receiving set. Both A and B pedestal pulses are coupled (during their respective periods) from the pedestal amplifier, V55, to the lower vertical deflection plate of V60, and to the automatic frequency-control circuits, as discussed. The accuracy of the delay readings obtained depends upon the accuracy with which the master and slave pulses can be superimposed (see figure 13-17, C).

The master and slave pulses may vary considerably in amplitude. An amplitude balance

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circuit, V61 (fig. 13-2) whose output is adjusted by the balance and gain controls, R249 and R256, provides a means of equalizing the master and slave pulse amplitudes in the A and B periods, respectively.

The sweep stages, V56, V57, and V58, produce the push-pull horizontal sweep voltages for the indicator tube, V60. The balancing d-c restorer, V59, blanks the indicator tube during the retrace periods.

The sweep function switch (fig. 13-1) may be moved to either of three positions to determine the type of pattern presented on the indicator tube. In SWEEP FUNCTION 1 position (see figure 13-17) the pedestal pulses are displayed on the A and B trace separated waveforms, respectively. In sweep function 2, the pedestals are not used for display purposes. In sweep function 3, superimposition of the traces is accomplished for matching of the two pulses (A and B), and trace separation is not used.

Figure 13-3 depicts the waveform components on the indicator tube when the sweep function switch (fig. 13-1) is in the SWEEP FUNCTION 1 position. The leading edge of the A pedestal is formed a short time after the reset of the A period. (The resetting action is accomplished by the reset thyratron, V37, of figure 13-2, as discussed.) This period is termed "A delay after reset." The end of this period (corresponding to the time of the leading edge of the A pedestal in figure 13-3) represents zero timing on the A and B period sweeps.

As the delay crank (fig. 13-1) is rotated clockwise, the B pedestal (fig. 13-3) moves to the right, and the delay counter reading (fig. 13-1) displays the time between reset of the B period and the leading edge of the slave pedestal, less the A delay. The time-difference counter reading is thus equal to the indicated time difference when the received master pulse is placed at the leading edge of the upper A pedestal, and the slave pulse is situated at the leading edge of the lower B pedestal.

In sweep function 2 (fig. 13-1) the sweep function switch, section S7D (fig. 13-2) permits the pedestal amplifier, V55, pedestal output to be applied to the sweep clamp stage, V56. The V56 output, in turn, drives the sweep generator, V57. At the same time, the pedestals are removed from the indicator display by S7C, and only the signals occurring on the pedestals during sweep function 1 are displayed by the indicator, V60.

For sweep function 3 of the sweep function switch, S7, the indicator trace displays the video signals which are present during the first 120 to 200 μs of the pedestal period. The trace separation stage, V62, is disabled by S7C, and the two traces are superimposed for matching of the received pulses.

ANTENNA COUPLER

The antenna coupler (fig. 13-1) permits matching of the antenna impedance to the input impedance of the receiver. The antenna coupler is shown schematically in figure 13-4. This unit consists of a 17-position rotary switch, S1; a tapped inductor, L1; a neon lamp, SD1; and a spark gap, E2.

When the rotary switch, S1, is adjusted for maximum signals on one of the four channels, the antenna and antenna lead-in reactance, and the section of the inductor, L1, in use, form a tuned circuit, which is broadly resonant over the band of frequencies that the equipment is designed to receive. The multiple taps on the inductance allow the circuit to be resonated with antennas having different characteristics.

The neon lamp, DS1, protects L1 and the receiver input components against burnout due to high voltages at the input of the antenna coupler. When the input voltage is greater than approximately 60 volts, DS1 will conduct. If the voltage is sufficiently high to cause it to burn out, spark gap E2, which has a higher breakdown, provides secondary protection.

R-F RECEIVER CIRCUITS

Attentuation and R-F Amplifier

The output of the antenna coupler (via J1 of figure 13-4) is applied to an attenuator switch (S2 of figure 13-5) through a coaxial cable. This switch is marked, local-dist, in figure 13-1, and has three positions corresponding to L, I, and D, respectively. The positions designate local, intermediate, and distant reception.

In the L (LOCAL) and I (INTERMEDIATE) positions of S2, T-type resistor network attenuators (fig. 13-5) are used to introduce attenuations of approximately 20 and 40 db, respectively. The proper impedance match for the interconnecting coaxial cable and the primary of T1 is maintained in all positions of the switch, S2. In the D (DISTANT) position, the input is applied directly to the primary of T1.

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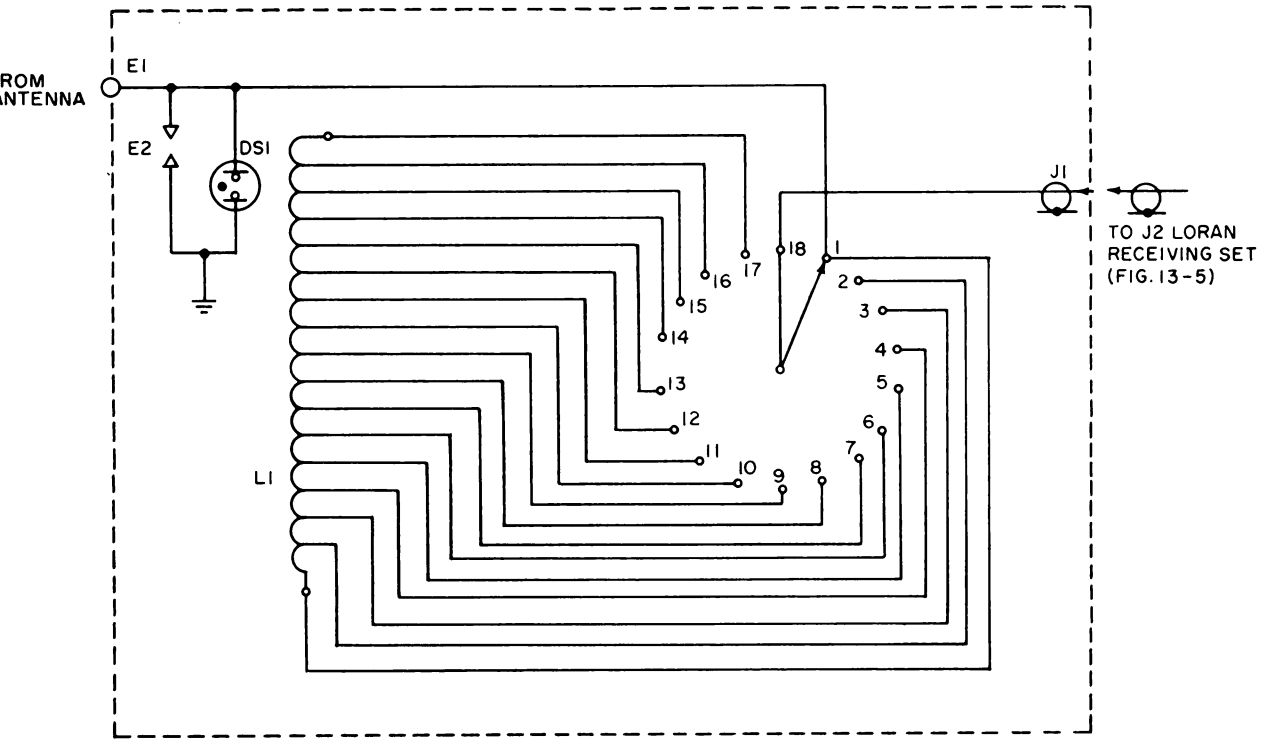

Figure 13-4.—Antenna coupler, schematic diagram.

Channel switch S3 when set for the desired hannel, selects a capacitor in section A of the witch to shunt the secondary of T1. The elected capacitor, in conjunction with C7, esonates the secondary at the desired fre- uency. Slug tuning is used to adjust the circuit o the highest frequency (1950 kc of channel 1), nd the trimmer capacitors for channels 2, 3, nd 4 (C4, C3, and C1, respectively) provide he necessary adjustment for each of the other hannels.

The r-f amplifier, V1, is a conventional cir- uit with the exception of an 1100-kc wave trap, Z2, in the cathode circuit. This circuit pro- vides degeneration at the intermediate fre- uency by inserting a high parallel resonant mpedance effectively in series with the signal ath to increase attenuation of signals at that requency in the V1 output.

Converter

The amplified output of V1 (fig. 13-5) is coupled to the signal grid (injector pin 7) of the converter, V2, by the double-slug, tuned f-f transformer, T2. Resistive loading at the primary of T2 by R12 is used to provide the desired bandpass characteristics. Sections B and C of the channel switch, S3, operate in the same way as section A to tune transformer T2 to the selected channel with the aid of the associated trimmer capacitors.

The cathode and first two grids of the converter, V2, are used in a Pierce-type oscillator circuit. The desired crystal is connected between pins 1 and 6 of V2 (through C30) by section D of the channel switch, S3. The crystal replaced the more common tank circuit. R14 and C26 develop grid-leak bias for the oscillator, and determine the amount of the feedback voltage across C26 to the pin 1 grid, relative to the total oscillator output signal. The output frequency of the converter is 1100 kc.

I-F Amplifiers

The converter (V2) output is amplified in the cascade i-f amplifiers, V3 and V4, before being applied to the detector. The three transformers, T3, T4, and T5, are double-tuned with resistive loading at both primary and secondary coils to provide the desired bandpass.

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The control grid of the r-f amplifier, V1, and both i-f amplifiers, V3 and V4, are returned to a control bias gain potentiometer in the balance circuit (fig. 13-18), which varies the grid bias to determine the gain of the stages. The operation of the balance circuit is treated later. The unbypassed cathode resistors of the r-f and i-f stages (fig. 13-5) provide a fixed minimum bias when the gain control is maximum (no grid bias applied to the stages). These resistors also produce a degenerative effect at the cathodes, which tends to prevent detuning of the r-f and i-f stages due to changes in grid bias.

Video Detector

The detector, V5A (fig. 13-5) is a conventional diode rectifier with a shunt R-C load consisting of R35 and C44. The V5A output is applied to the grid of the first video amplifier tube, V6A.

Video Amplifiers

The first video amplifier, V6A, also serves as a limiter when the negative input voltage exceeds the cutoff value of the tube. The positive-going output of V6A is fed to the second video amplifier, V6B, through the antijam switch, S4.

The antijam switch, S4, when in the IN position (as shown) couples the positive output of V6A to the grid of V6B through the high-pass filter consisting of C46, C47, and R41. This filter reduces the effects of jamming by removing the lower-frequency components of the video signal.

The clipper stage, V5B, removes negative peaks from the V6B input. The negative pulses are derived from differentiation in the short time constant circuit comprising C46, C47, and R41.

When the antijam switch is in the OUT position, the output of V6A is coupled through the 6-8 contacts of S4, through C48 and R38, and through the 2-12 contacts of S4, to the V6B grid. C48 and R38 form a voltage divider with R41 to give an attenuation, which is equal to the loss obtained when the high-pass filter is used. Thus, the input to V6B for a given signal level remains the same whether the switch, S4, is in the IN or OUT position, and no adjustment of the front panel gain controls is necessary.

The video gain control, R41 (a screwdriver adjustment) sets the level of the video output signal in both positions of the antijam switch. The video signal from V6A is amplified by V6B and coupled through C54, through a self-check switch, S10 (discussed later), and through R61 to the video amplifier, V7. The video is also applied through R60 to the grid of another video amplifier via pin E of J5. The purpose of the J5 output will be considered later.

Video amplifier V7 is essentially a clipper and amplifier stage. It should be recalled that V5B operates to remove negative portions of the video output of V6A. However, it is quite possible that some of the negative portions of the video, which are generated by differentiation at the V6B grid as discussed, will be developed across R41, and therefore amplified in V6B. These negative voltages at the V6B grid will appear as positive voltages at the V6B plate, which are also coupled by C54 to the V7 grid, causing the V7 grid to draw a current through grid limiting resistor R61. This action develops a negative voltage at the V7 grid, and therefore limits positive alternations of the grid input.

Negative alternations at the V7 grid (from V6B via C54) are amplified by V7, and appear as positive-going pulses at the V7 plate. These positive video pulses are coupled by C55 to the upper deflection plate of the cathode-ray tube, V60.

TIMER AND INDICATOR

To this point in the discussion, the loran signal input has been selected (by the antenna, r-f amplifiers and converter stages), amplified (in the i-f stages), detected (by video detector), and fed through stages of video amplification and clipping to the upper vertical deflection plate of the CRT. To hold the loran signal stationary on the screen of the indicator tube while readings are being taken, it is necessary to employ frequency control circuits to synchronize the repetition rate of the loran receiving set with the loran signals from the transmitting pair. This action is accomplished in the timer and indicator chassis (fig. 13-1).

Oscillator and Reactance Circuits

The time base for the synchronization system in the receiver is an 80-kc sine wave oscillator, V17 (fig. 13-6). The oscillator, V17, is crystal controlled by Y5 (at the normal frequency of

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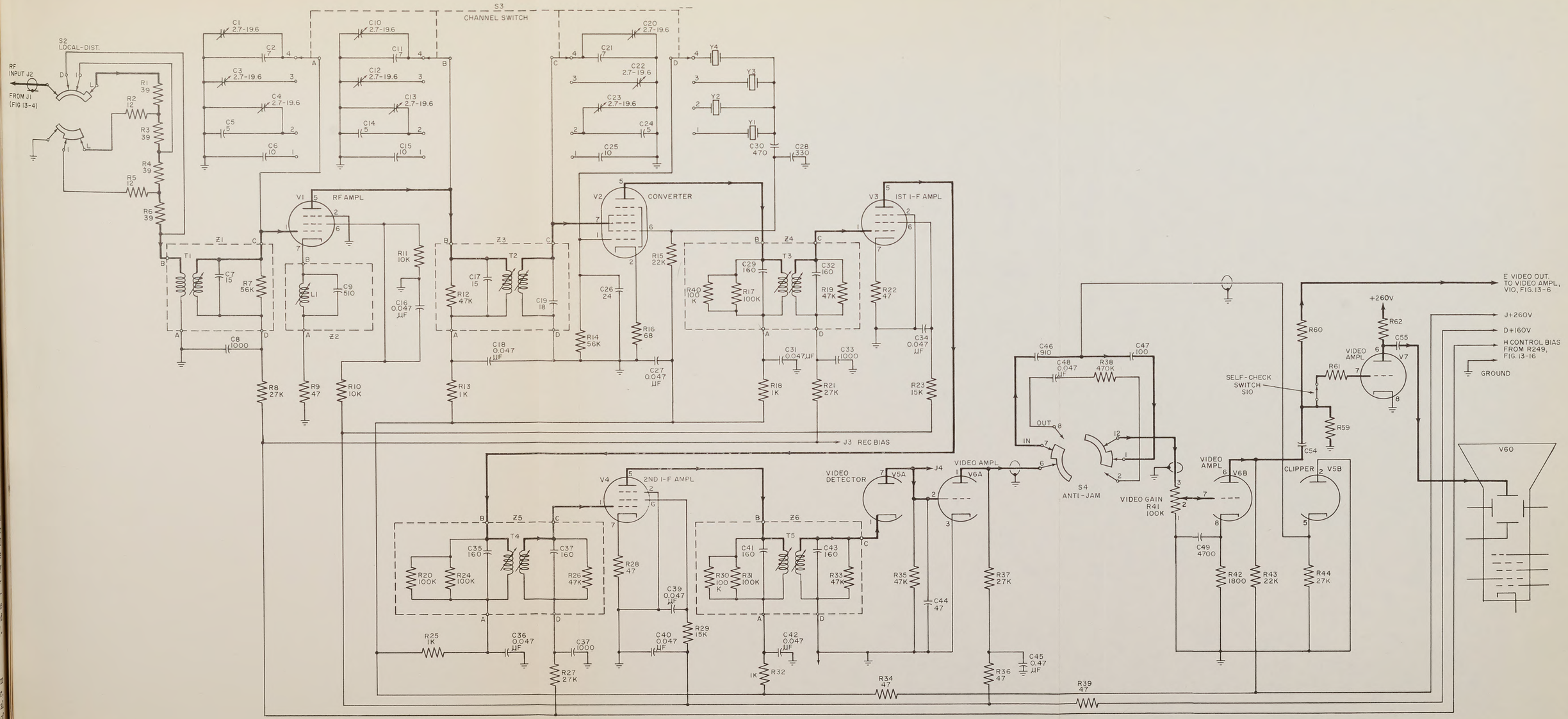

Figure 13-5.—R-f section, schematic diagram.

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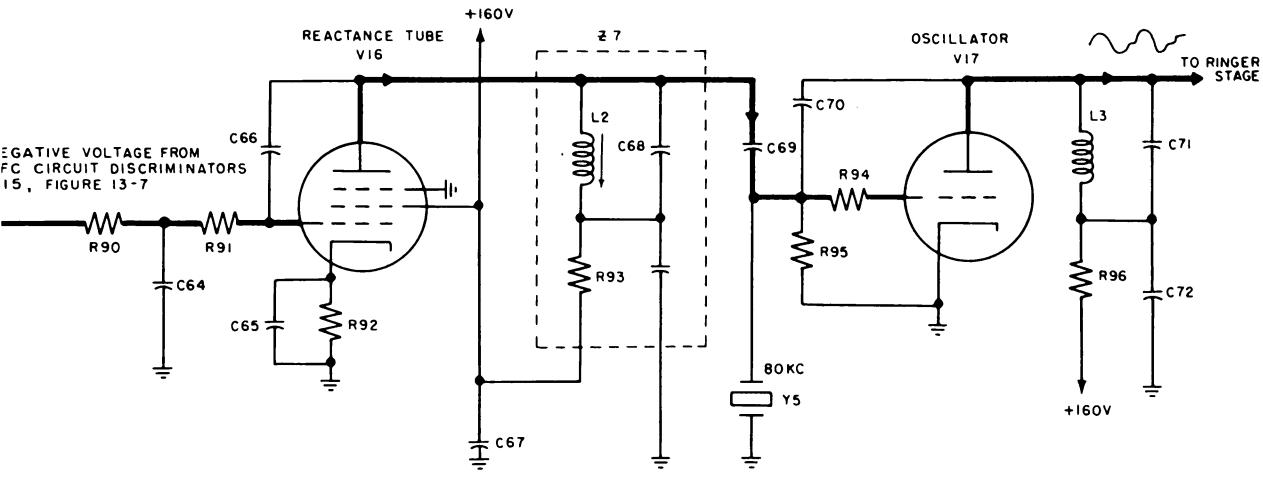

Figure 13-6.—Frequency control circuits (oscillator and reactance tube).

0 kc), and is essentially a tuned-plate, tuned-rid oscillator. The 80-kc crystal is connected n series with parasitic suppresser R94 at the rid of V17 to prevent oscillations at frequencies ot controlled by the oscillator. Feedback is rovided by the interelectrode capacity of V17 nd is supplemented by C70, which is connected etween the plate and grid elements of V17. rid-leak bias is obtained by the discharge of he capacity of the crystal holder of Y5 through 95. C69 is a blocking capacitor, which isolates he oscillator (V17) circuit from the d-c plate oltage on the reactance tube, V16.

The reactance tube, V16 (fig. 13-6) receives t its grid a small portion of the 80-kc voltage ppearing across the crystal through the network onsisting of C64, R91, C66, and C69. The V16 rid is held negative to the cathode by a negative oltage received from the AFC circuit (discussed ater) supplemented by cathode bias developed at he V16 cathode by R92 and C65.

The voltage at the grid of the reactance tube, 16, leads the voltage at the grid of the scillator, V17, by approximately 90° because he reactance of capacitor C66 between the 16 grid and plate is much larger than the esistance of R91 between grid and ground, nd the reactances of C67 and C69 are relatively mall. The plate load of the reactance tube, Z7, s resistive when resonated at 80 kc. Thus, the late load current is in phase with the grid oltage and therefore leads the plate voltage by 0° This stage therefore acts like a variable apacitor coupled by C69 in parallel with the 0-kc crystal.

The value of the apparent capacity across the crystal, Y5, depends upon the gain of the reactance tube, V16, which in turn depends upon the V16 bias. The value of bias is controlled (to an extent) by a negative voltage from the AFC circuit. The range of frequency control as a result of the action of the reactance tube, V16, is small (less than ±30 cps), but is sufficient to compensate for differences in pulse repetition rate between the transmitting and receiving systems.

Frequency Control Circuits

As stated earlier, the basic timing system of the receiving set is controlled by the 80-kc crystal oscillator, V17. However, to hold the received pulse stationary on the screen of the indicator tube, the timing base must be synchronized to an exact multiple of the reception rate of the received signals.

Both AFC (Automatic Frequency Control) and manual drift circuits are available for use in synchronizing the receiving set with the received signals. In both cases, the control is accomplished by variation of the grid bias applied to the reactance tube, V16. The AFC circuit automatically changes the grid bias of V16, which changes the constants of the oscillator, V17, sufficiently to lock in the receiving set repetition rate at the proper relation to that of the transmitting group. Adjusting the drift control on the receiver front panel (fig. 13-1) will manually set the grid bias to control the reactance tube, V16 (fig. 13-6).

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The automatic frequency control circuits are shown in figure 13-7. V8 and V9 are one-shot multivibrators referred to as the A and B multivibrators, respectively. The one-shot cycle is described below. In the quiescent condition, V8B conducts, and V8A is cut off. The positive pulse on the V8A grid causes V8A to conduct. The resulting decrease in V8A plate voltage is coupled through C52 to lower the V8B grid below cutoff. The V8B plate voltage rises sharply as V8B plate current is cut off. C52 discharges through R51 (for 80 μs) to raise the V8B grid voltage above cutoff at which time V8B again conducts. The accompanying sudden increase in voltage across R52 cuts off V8A, and the one-shot cycle is completed.

The A multivibrator, V8, receives a positive rectangular trigger pulse from the pedestal amplifier. (The origin and purpose of the pedestal pulses are treated later.) The control grid of V8B is returned to the +160-volt supply through R51, and is therefore the normally conducting section of V8. The positive rectangular output of the pedestal circuit is coupled through isolating resistor R46 and a differentiator circuit consisting of C50 and R45. The positive pulse resulting from the differentiation of the input across R45 triggers V8A into conduction to initiate the A multivibrator cycle.

The A multivibrator, V8, furnishes a positive rectangular pulse from V8B through R50, to the suppressor grid of the A gated amplifier tube, V11, and a negative rectangular pulse from V8A, through C51 to the B multivibrator, V9. The duration of the output of each section of V8 is approximately 80 μs. The pulse furnished to the grid of the B multivibrator, V9, is differentiated in the network comprising C51 and R47 to form a sharp negative pulse. This pulse occurs at the same time as the leading edge of the input pulse to V8A. The operation of the B multivibrator, V9, is identical to that for the A multivibrator, V8.

Since a positive pulse from V8A is required to trigger V9A, the V9A conduction occurs 80 μs later (in time) than V8A when the V8A plate output (shown in dotted lines) is positive going. The positive rectangular output of V9B is applied through R57 (80 μs later than the pulse from V8B) to the suppressor grid of the B gated amplifier, V12, and is also approximately 80 μs in duration.

The A and B gated amplifiers, V11 and V12, respectively, are normally held beyond cutoff by the negative bias (from the -180-volt supply) applied to the suppressor grids of V11 and V12 through R63 and R58, respectively. The suppressor grid of V11 is connected to the junction of R63 and R50. R63 connects to the -108-volt supply, while R50 is connected to the plate of V8B. Before the A multivibrator, V8, is triggered (at V8A) the plate voltage at V8B is approximately 45 volts, and the voltage divider comprising R50 and R63 applies about -36 volts to the suppressor grid of V11.

When V8A receives its positive trigger pulse, this tube section conducts, and the accompanying drop in the V8A plate potential is fed through C52 to the grid of V8B. V8B is cut off as a result of the negative-going grid input from V8A, and the V8B plate voltage rises toward the +160-volt supply potential. Likewise, the suppressor grid potential at the V11 grid (from the R50-R63 junction) rises from -36 volts to approximately zero volts.

Note at this time that V11 can also receive the positive video pulses from V10 at its control grid if the AFC switch, S5, is in the ON position, as shown. V10 receives its video input from V6B via pin E of J5 (fig. 13-5), and functions the same as V7B (in the same figure) to limit and amplify the video signals.

During the period that the positive trigger from V8B prevails at the V11 suppressor grid (duration approximately 80 μs), a coincident firing action occurs in V11 if the positive video output of V10, coupled by C64, is simultaneously applied at the V11 control grid.

Video signals may or may not be applied to the V11 grid from V10, depending on whether or not the system is receiving an antenna input signal. If no signal is being received, about -7 volts bias is established between the control grid and cathode of V11 (and V12) from the junction of R65 and R66. These resistors are connected between the -108-volt supply and ground.

The operation of V11 and V12 is identical. The control grid of both tubes is coupled to the plate of V10 by C64. The video amplifier (V10) output is differentiated by C64, R64, and R65 so that, regardless of the received video signal pulse width, the input of the A and B gated amplifiers, V11 and V12, is held to approximately the same width as that of the gated pulses from the A and B multivibrators.

When the (V10) video amplifier output occurs during the first 80-μs interval after the trigger pulse from the pedestal amplifier (corresponding in time to the leading edge of the rectangular pulse to V11 from V8B), the video signal will

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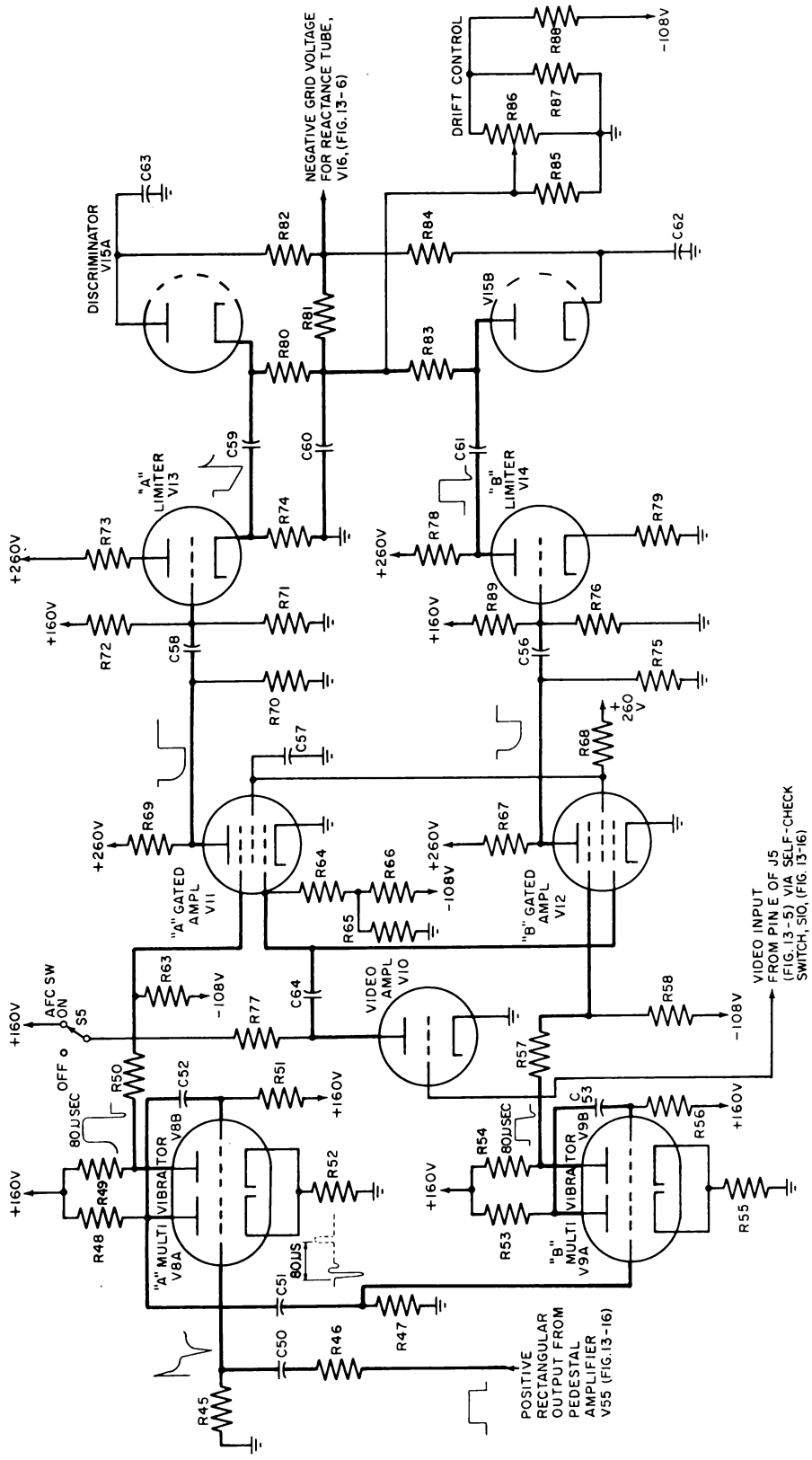

Figure 13-7.—Frequency control circuits (A and B gating circuits and discriminator).

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be amplified only by the A gated amplifier, V11. Conversely, if the video signal occurs between 80 and 160 μs after the leading edge of the pedestal pulse (corresponding in time to the trigger pulse to V12 from V9B), the video will be amplified only by the B gated amplifier, V12. Both gated amplifiers, V11 and V12, will conduct when the video from V10 straddles the two gated periods.

The A gated output of V11 is coupled by C58 to the A limiter, V13. The B gated output of V12 is coupled by C56 to the control grid of the B limiter, V14. Both A and B limiters (V13 and V14, respectively) are normally conducting as a result of positive voltage at the grid of each of the tubes developed across R71 and R76, respectively. These resistors are connected through R72 and R89, respectively, to the +160-volt supply.

The A limiter, V13, clips the positive peaks of its input signal from V11, but amplifies negative portions of the signals. The output at the V13 cathode is negative (less positive) pulses, which are coupled by C59 to the discriminator, V15A.

The B limiter, V14, also clips positive peaks of the video input. Negative portions of the video appearing at the V14 grid are developed at the V14 plate as positive pulses, which are coupled by C61 to the discriminator, V15B.

Discriminator

The discriminator, V15, utilizes the gated outputs of the A and B circuits comprising V8 through V14 to produce a d-c output voltage, the polarity of the discriminator output depending upon whether the received signals are faster or slower than the receiving set repetition rate. If the receiving set repetition rate is in exact synchronism with the received signal, the input pulses to the discriminator from the A and B circuits will produce equal and opposite voltages, and the discriminator resultant output is zero.

The discriminator output at the R81-R82 junction is applied to the grid of the reactance tube, V16 (fig. 13-6), causing the V16 output to the oscillator, V17, to change as the voltage of the V16 grid changes. Accordingly, the oscillator (V17) output changes until it is synchronized to an exact multiple of the receiving set repetition rate, and the patterns being observed on the indicator will remain stationary.

The operation of the discriminator circuit is explained with the use of the simplified diagram of figure 13-8. The diagram shows the conducting path for each of the discriminator diodes V15A and V15B, with the path of V15A shown by the solid arrows, and the path for V15B indicated by the dotted arrows. If the video signal from V10 arrives at the A and B gated amplifiers only in one of the gate periods (either A or B), the receiving set repetition rate and the signals being received are not in synchronization.

When the video pulse from V10 (fig. 13-7) is present at the V11 control grid during the first 80-μs interval after the arrival of the pedestal trigger pulse to the A multivibrator, V8A, the resulting negative-going output of the A limiter, V13, as discussed, is coupled by C59 to the discriminator (V15A) cathode. Thus, V15A, (fig. 13-8) will conduct for the duration of the pulse (80 μs) to charge C63 negative to ground. The charge developed across C60 as a result of the input pulse is small due to the relatively large size of R80 in the C60 charge path.

After the termination (trailing edge) of the input pulse, C63, begins discharging (fig. 13-8, B) through R82, R81 and C60. The resulting potential at the R81 and R82 junction is negative to ground during the charge and discharge of C63. This negative potential is applied to the control grid of the reactance tube, V16 (fig. 13-6), and aids the existing negative potential at the V16 grid applied from the arm of the drift control, R86 (fig. 13-7). The reactance tube, V16, will affect the oscillator, V17, in a manner that causes the oscillator frequency to move in a direction that opposes the drift of the video pulse into the A gate period. The oscillator frequency is continuously corrected until the video pulses straddle the A and B gate periods. When this condition is reached, the system is synchronized to the received signals.

If the video pulse from V10 (fig. 13-7) arrives during the second 80-μs interval after the arrival of the trigger pulse at V8A, the B limiter, V14, will supply a positive pulse through C61 to the V15B discriminator plate (fig. 13-8). The resulting conduction of V15B charges C62 positive to ground, as shown in dotted lines in figure 13-8, A.

At the trailing edge of the video pulse applied to the V15B plate, C62 begins to discharge (fig. 13-8,C) through C60, R81, and R84. The potential now being developed across R81 is

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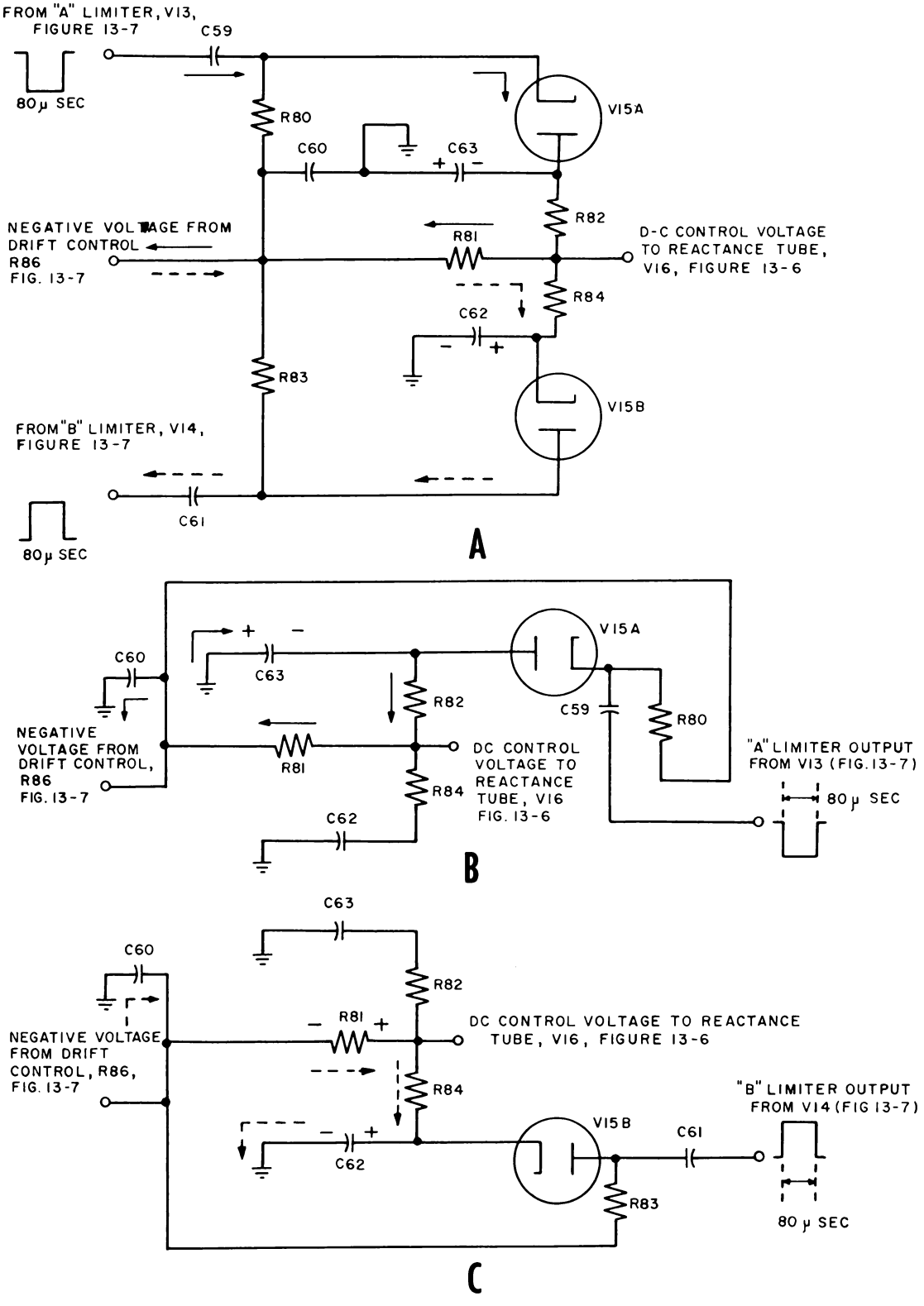

Figure 13-8.—Discriminator circuit operation, simplified schematic diagram.

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positive-going, and thereby reduces the negative voltage at the reactance tube (V16) grid (fig. 13-6). The reactance tube will affect the oscillator frequency in the opposite direction of that which resulted for the V15A output to oppose the drift of the video pulse into the B gate period. When the received video pulse straddles the A and B gate periods, the receiving system is synchronized with the received signals.

When the AFC switch, S5 (fig. 13-7) is in the ON position, the discriminator operates in conjunction with the drift control, R86, to determine the bias applied to the reactance tube (V16) grid, as discussed. Normally, when the AFC circuits have locked in on the received signal, the drift control, R86, may be varied throughout a large part of its range without affecting the AFC circuit because the voltage developed across R81 will compensate for changes in the voltage applied by the drift control. However, if the AFC circuit should lose control due to a sudden noise pulse or interfering signal, readjustment of the drift control setting would stop the signal movement.

With the AFC switch, S5, in the OFF position, the plate voltage of the video amplifier, V10, is removed. Thus, the signal to the discriminator is removed, and the drift control governs the bias on the reactance tube, V16.

Pulse Forming Circuits

Before studying the operating details of the pulse forming circuits, consider the following general characteristics of the circuit. Assume that the receiving system is synchronized with the received signals as a result of the discriminator action. The receiving set must generate the pulse recurrence rate to which the system is synchronized. To perform this function a group of 12 frequency counter stages (discussed later) is used in the receiving set to obtain subharmonics of the 80-kc crystal oscillator frequency produced by V17 (fig. 13-6), as discussed. Actually, the subharmonics produced by the counter stages are used for three major purposes: to obtain the receiving set pulse repetition rate as selected by the PRR (Pulse Repetition Rate) switch on the receiver front panel, to produce the trigger pulses used to initiate the master and slave pedestals, and to provide two square waves equal in duration to the recurrence interval for the trace separation and amplitude balance circuits.

The 12 counter stages are multivibrators, which are driven to produce an output square wave. The first 11 multivibrator stages are connected in cascade so that each stage in the chain supplies the necessary driving pulse for the following stage. The output square wave from each multivibrator has a period twice that of the recurrence rate of the pulses that drive the multivibrator. Only one driving pulse is obtained from each half of the square-wave cycle (for a given counter), and the output frequency of the following counter stage is thus one-half that of the preceding one.

Outputs from the first 11 counter stages produce the square waves used to obtain the receiving set repetition rate. In addition, the first 11 counter stages produce outputs, which are used in the variable delay circuits (discussed later) to produce a variable B pedestal pulse. The twelfth counter stage is driven by a reset pulse to produce a square wave output, which is equal in duration to the receiving set pulse recurrence interval.

The ringer and clipper stages, V18 and V19 (fig. 13-9) are included in the pulse-forming circuits to produce the sharp, negative driving pulses that are required to drive the counter chain. The ringer stage, V18, receives the output of the oscillator (V17 of figure 13-6) through C73. The cathode of the ringer, V18 (fig. 13-9) is grounded, and the oscillator (V17) output drives the ringer (V18) grid positive on positive peaks of the oscillator output. The V18 grid draws a current through the low cathode-to-grid resistance of V18 to charge C73 negative to ground. Due to the long time constant of the coupling components, C73 and R97, as compared to the period of the 80-kc oscillator input, the discharge of C73 through R97 holds the V18 grid beyond cutoff except on input positive peaks. The action is that of grid-leak bias at the V18 grid.

L4, in the plate circuit of the ringer tube, V18, is resonated by its distributed capacitance, Cd, and by the capacitances to ground of its associated components. During positive peaks of oscillator V17 (fig. 13-6) output to the grid of V18 (fig. 13-9), the ringer tube conducts, and a negative-going pulse is applied to L4. The exciting pulse from the V18 plate causes L4 to ring (oscillate) and produce a train of oscillations at the resonant frequency of the circuit. The L4 output is coupled by C76 to the grid of clipper stage V19. C76 and R99 develop grid-leak bias at the V19 grid.

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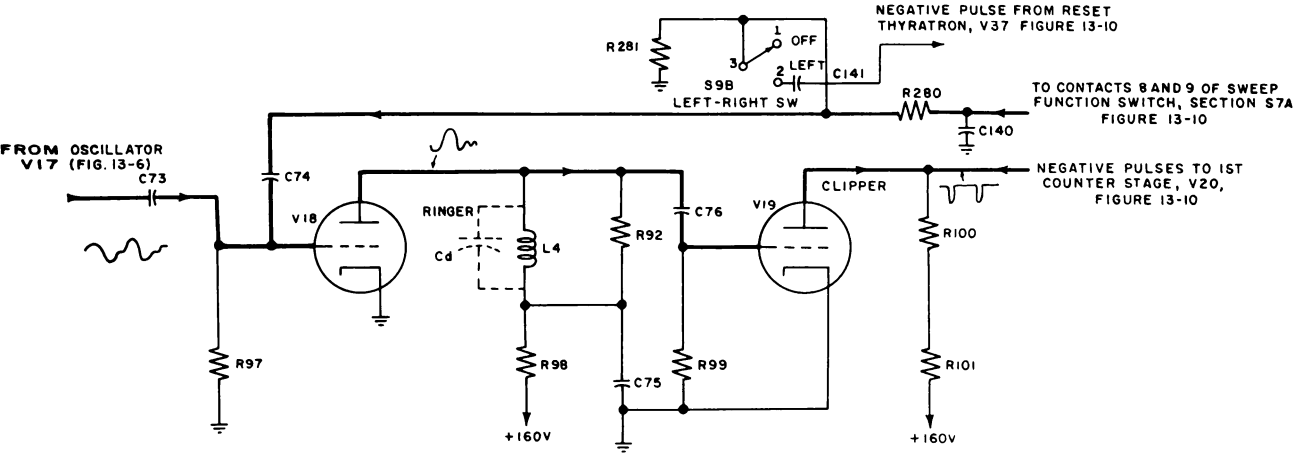

Figure 13-9.—Pulse-forming circuits (ringer and clipper).

R92 is connected across L4 to provide a damping action to the L4 oscillations, so that only the positive first half-cycle of the damped train is of sufficient amplitude to overcome the grid-leak bias at the clipper tube (V19) grid. The output of the clipper stage, V19, is sharp, negative pulses, which are coupled to the first counter stage, V20 (fig. 13-10).

The counter stages, V20 through V31 (fig. 13-10) are flip-flop (Eccles-Jordan) multivibrators. The cascade connection of the 12 multivibrators allows the output pulse of one counter stage to serve as the input trigger pulse for the following stage, as mentioned earlier. Thus, V20B supplies the input trigger to V21A. V21B triggers V22A, and so on.

Since the operation of each of the multivibrators is the same, only one of the circuits will be explained with the aid of figure 13-11. The first counter, V20 (fig. 13-10) is represented in the illustration without regard to the action of the L-R switch, S9. The purpose of this switch and the input applied through the S9 contacts will be explained later.

The negative trigger pulses from the clipper stage, V19 (fig. 13-9) is applied through C78 (fig. 13-11) to the junction of two diodes, CR1 and CR2, which are connected back-to-back between the two plates of the counter stage, V20. The potential at the R108-CR1 and R110-CR2 junctions depends upon the conducting or nonconducting condition of V20A and V20B, respectively.

Assume that V20A is conducting and V20B is nonconducting. The plate voltage of V20A will be low, and the voltage across R105 and R106 is low. Likewise, only a relatively small positive potential will be applied to the grid of the nonconducting tube, V20B, from the R105-R106 junction. The cathode voltage developed across R118, R117, and R116 will be sufficient to maintain V20B beyond cutoff.

Conversely, the voltage at the plate of the nonconducting tube, V20B, is high, and the drop across R112 and R113 is correspondingly larger than that developed by R105 and R106. Thus, a larger positive potential is applied to the V20A grid from the R112-R113 junction, which aids the conduction of this tube section to maintain V20A at saturation.

When the negative trigger pulse from V19 (fig. 13-9) is applied at the junction of R109 and the two diodes, CR1 and CR2 (fig. 13-11), the diode connected to the "off" plate will have the larger voltage difference placed across its terminals, and will therefore conduct a heavy current. The conduction path is through the low conducting resistance of V19 (fig. 13-9), and the low charging impedance of C78 (fig. 13-11), and the low conducting resistance of CR2, through R110, and through the parallel combination of R114 and R115 to the B supply. Note that the current through R110 and R114 and R115 (in parallel) is very much increased, with a corresponding increase in voltage across these resistors. Also note that the resistance from the R110-CR2 junction to ground is decreased (by the low impedance parallel path comprising V19, C78, and CR2), and the voltage from the R110-CR2 junction to ground decreases a corresponding amount. Thus, the potential at the R112-R113 junction is also decreased, and the

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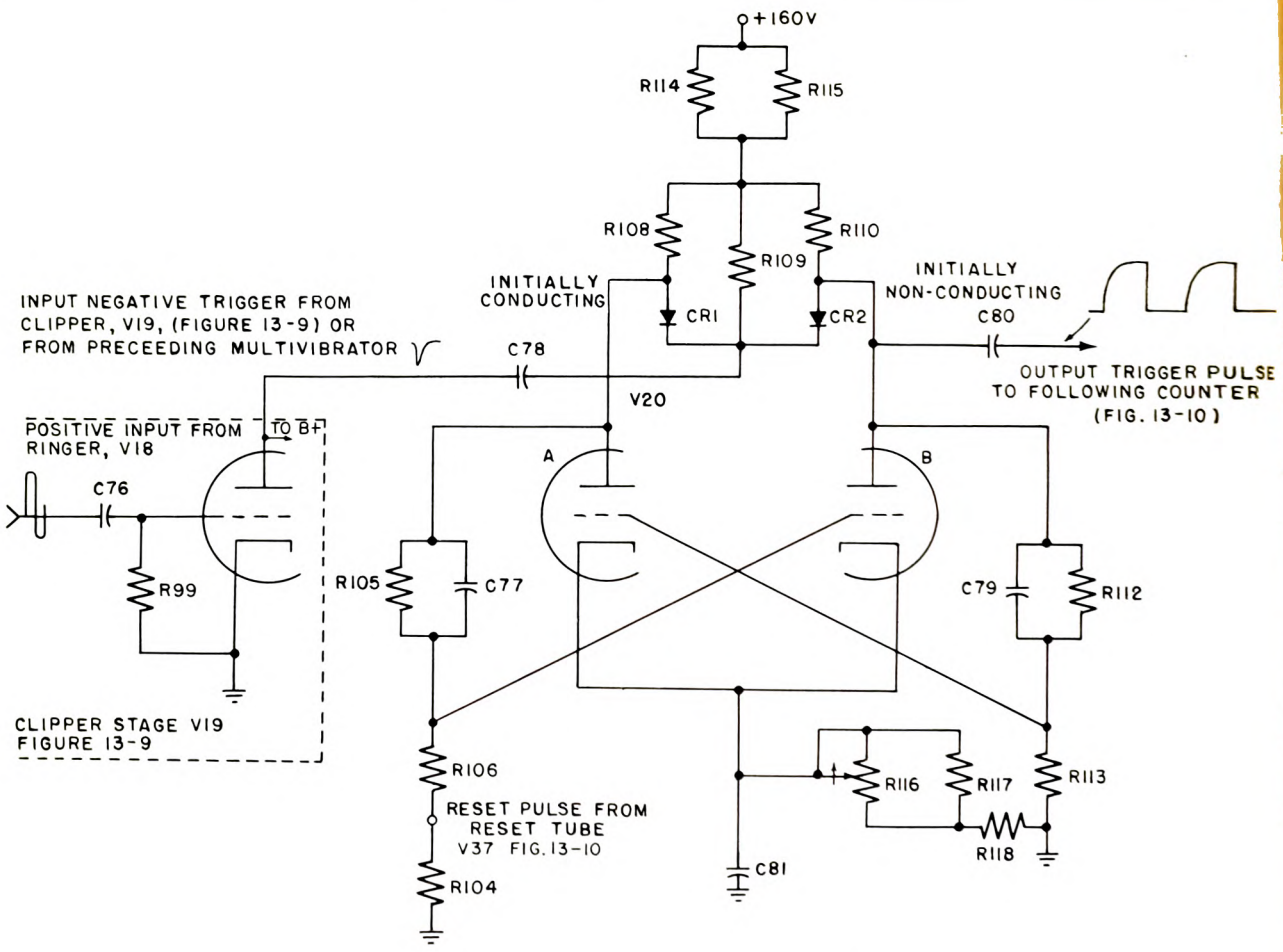

Figure 13-11.—Representative counter circuit.

positive potential applied to the V20A grid decreases. This action moves the conducting point of V20A below saturation.

The plate voltage rise of V20A (when the tube current is decreased from saturation current) causes the voltage across the R105-R106 divider to become more positive. The more positive potential at the R105-R106 junction decreases the grid-cathode voltage of V20B to allow this tube section to begin conducting. The corresponding drop in plate voltage of V20B acts as regenerative feedback to the V20A grid from the R112-R113 junction. Thus, V20A goes to cutoff as V20B goes to saturation.

Due to the action of the diodes, the trigger is applied only to the nonconducting stage. The regenerative action (as discussed) originating from the effect of the input pulse on the nonconducting tube causes the generation of a square wave having a fast-rise and fast-fall time. The counter stage output (V20 in this case) is taken from the B section of the tube, and applied through C80 to the following counter stage (fig. 13-10).

The cathodes of all the counter stages (V20 through V30) are returned through the counter bias (CTR BIAS) network consisting of R116, R117, and R118 to ground. R116 can be adjusted to set the counter bias at the proper level.

The pulse repetition rate coincidence stages, V32 through V36A, utilize the various outputs of the counter chain, V20 through V31, to derive a particular pulse repetition rate for the receiving system chosen by the PRR switch on the receiving set front panel. Because each counter stage must be triggered twice by a negative pulse before it produces a trigger pulse for the following stage, the point (in time) at which any

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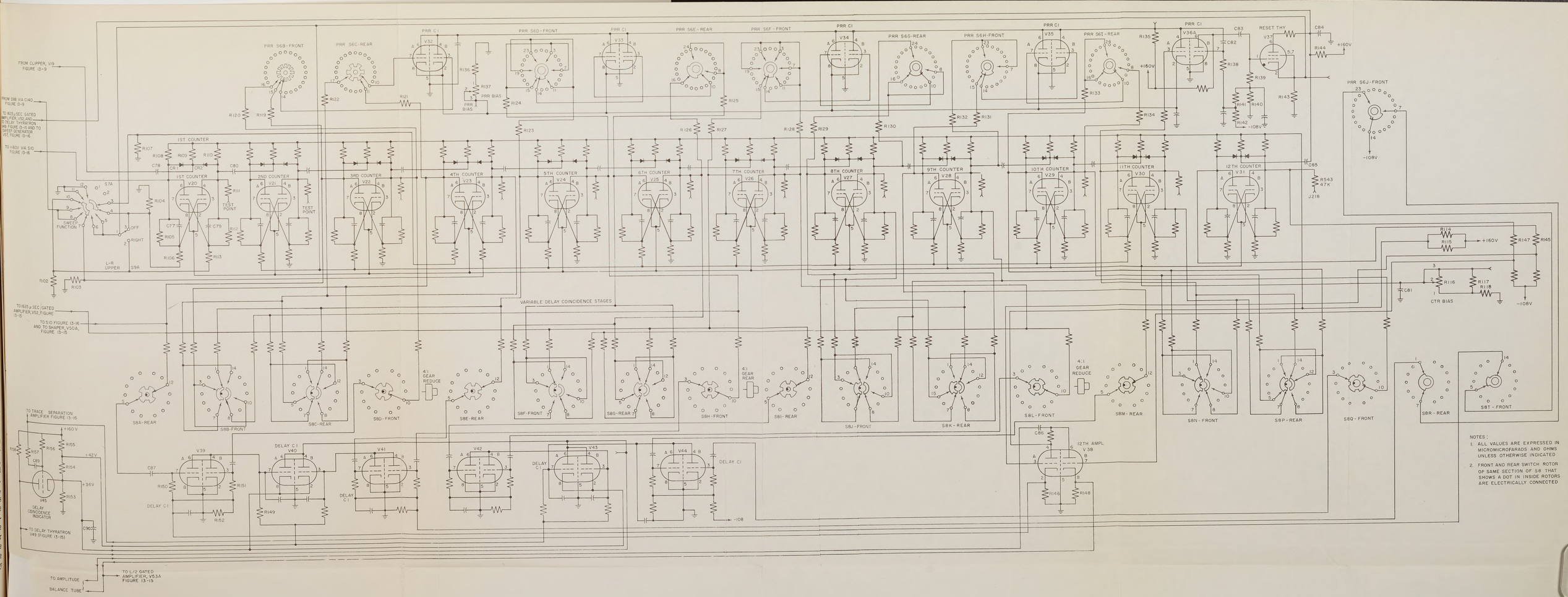

Figure 13-10.—Counter, pulse repetition rate,

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tube causes the generation of a following stage, the point (in time) at which any

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tube switches can be represented by a given count of driving pulses applied to the input of the counter chain.

By combining the voltage values of the output square waves present at the various counter plates, it is possible to determine the point at which a given count is reached. The count will correspond to a selected pulse repetition rate.

Let us analyze this concept with the aid of the plate waveforms of V28A, V29A, and V30A shown in figure 13-12. Each of the three plate outputs of V28A, V29A, and V30A of figure 13-10 will be negative going only at counts between 1792 and 2048 on the scale shown (fig. 13-12). This coincident occurrence of all negative voltages at the chosen plates correspond in time to a given count selected by the PRR switch.

If the B section output of V29 shown on dotted lines were substituted in place of the A section output, the simultaneous, negative indication would be obtained only at counts from 1270 to 1536. This corresponds in time to a different setting of the PRR switch.

It will be shown later that the negative voltage obtained for a selected count (as determined by the PRR switch) controls the resetting of all the counter stages to start a new receiving cycle, thereby determining the pulse repetition rate of the receiving system.

It is possible to obtain a negative output from the counter stages during any of the intervals plotted on the bottom scale (if the proper counter plates are chosen). If the outputs of all the binary plates were properly combined, it would also be possible to secure a unique negative indication for every count from 0 to 2048. Since only 24 different repetition rates are used by the various transmitting groups (see table 13-1), it is not necessary to use all of the outputs of the first 11 counter stages.

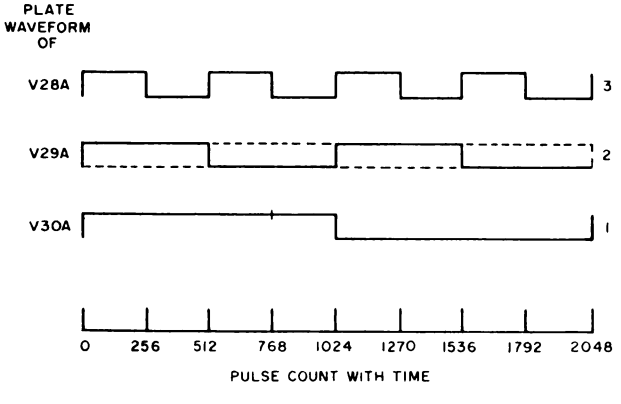

Figure 13-12.—Negative coincidence indication from combined plate waveforms.

The plates of the third through the tenth stages of the counter chain (V22 through V29 of figure 13-10) are direct-coupled through their respective isolating resistors, R119 through R134, to sections B through I of the PRR switch, S6. The switch selects the proper counter plates to produce the negative output that corresponds to a predetermined count. The chosen plate output is coupled through an isolating resistor, and through the chosen switch contacts to the grid of one of the amplifying tubes, V32 through V35.

The pulse count (fig. 13-12) represents the number of pulses in time required to produce the effect indicated for either of the tubes V28A, V29A, and V30A. Only counts above 1024 are used. The A plate of V30 is negative for the entire period from 1024 to 2048. The negative pulses from V30A are always used in determining the pulse repetition rate of the receiving set, and are applied through R134 (fig. 13-10) to the V36A control grid.

The plates of V32 through V36A are connected together and returned through a common load resistor, R135, to the +160-volt supply. Cathode bias for V32 through V36A is obtained from common cathode resistors R136 and R137. PRR bias control R137 is variable to permit the bias to the tubes to be adjusted.

Recall that the input (grid) voltage of any one of the tubes, V32 to V35, may be positive or negative, depending upon the potential applied from the counter plate to which the grid of the tube is connected through a section of the PRR switch, S6. The V36A input from V30A may also be either positive or negative. When the input pulse from a counter to any section of the coincidence indicator chain comprising V32 through V36A is positive, that section will draw saturation current. When the counter input to a particular section of any one of the tubes (V32 through V36A) is negative, that section will be held beyond cutoff for the duration of the counter input. V32 through V36A are paralleled, with the common plate load resistor, R135, connected to the +160-volt B supply.

The d-c conducting resistance of any one of the tubes, V32 through V36A (either singly or in parallel) is low as compared to the series resistance of R135 (100 kilohms). Thus, the voltage drop across R135 does not vary greatly

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with one section of the coincidence indicator chain tubes (V32 through V36A) conducting, or with all sections of the tubes conducting. The only sharp changes in voltage appear when all of the tubes, V32 through V36A, are cut off (corresponding to high d-c resistance of the coincidence indicators), and at the point where any one of the tubes begins conducting (corresponding to a drop from a very high to a very low d-c plate resistance of the tubes).

Since an output is produced from the coincidence indicator chain during the time that all sections of V32 through V36A are cut off, the width of the output cannot be greater than the width of the negative half of the smallest square wave applied to the chain. The first counter output pulse to the coincidence indicator chain is applied from V22A (or V22B) through S6B to one of the grids of V32. One of the V22 outputs will always be applied to V32, depending on the position of S6B. The output frequency of the third counter V22, is 10 kc, and the width of the negative pulse is 50 μs (one-half the total pulse duration). This represents the maximum time that all of the tubes in the coincidence indicator chain (V32 through V36A) can be held cutoff, and likewise sets the output pulse width of the coincidence indicator.

Changes in the plate to ground voltage of V32 through V36A are coupled by C82 to the control grid of the clipper amplifier stage, V36B. The V36B grid is normally held at about -34 volts negative to the cathode, and V36B is nonconducting. Positive inputs at the V36B grid cause an output at the V36B plate. Negative input voltages are not amplified by V36B because negative inputs drive the V36B grid further into cutoff.

The negative V36B output is differentiated in the short, time-constant circuit comprising C83, R139, and R140. The positive pulse resulting from the differentiation is applied to the grid of a reset thyratron, V37.

The reset thyratron, V37, is normally held beyond cutoff by a fixed bias voltage from the -108-volt supply at the R140-R141 junction. The positive differentialed output of V36B (when the coincidence chain comes back into conduction) triggers V37 into conduction. The resulting positive output at the V37 cathode across R143 is a very sharp positive pulse. The point (in time) at which the reset pulse occurs corresponds to 50 μsec after the time during which the simultaneous negative voltages from the counter plates (as discussed) arrives at the grids of V32 through V36A. Thus, the recurrence rate of the reset pulse at the V37 cathode is determined by the selection of the counter plates by the PRR switch, S6, to supply the simultaneous negative pulses to the grids of all the coincidence indicator chain tubes, V32 through V36A. The output reset pulse from V37 is fed from the V37 cathode to the control grids of the B sections of V21, V22, and V24 through V30 (through their respective grid resistors). The reset pulse is also applied to the first counter, V20, and to the fourth counter, V24, if the L-R switch, S9, is in the OFF position.

The reset pulse causes a switching of all the conducting tube sections in the counter chain, and thereby resets the chain. When the point (in time) is again reached where all of the counter outputs applied to the coincidence indicator chain are negative, another reset pulse will be produced 50 μ sec later and will likewise again reset the counter chain. Thus, the pulse repetition rate (frequency) at which this action occurs is determined by the time required for the counter chain to reach the condition of all negative outputs to the coincidence indicator plus 50 μ sec. This, in turn, corresponds to a given setting of the PRR switch, S6.

The positive reset pulse from the V37 cathode will appear at the B grids of the second, third, and fifth through eleventh counter stages, as a result of a permanent connection. With the L-R switch, S9A, in the position shown, the positive reset pulse from the cathode of V37 is also applied through contacts 1 and 3 of S9A, and through R106 to the B grid of the first counter stage, V20, and from contact 3 of S9A through contacts 4 and 3 of S7A to the B grid of the fourth counter, V23. Thus, all counter stages are reset on the same side at a rate determined by the PRR switch, S6.

The trailing edge of plate output of the reset thyratron, V37, is shaped by the charge of C84 toward the +160-volt supply through R144, when V37 is cut off. Thus, the trailing edge of the V37 plate waveform rises exponentially. The various uses of the V37 plate output will be considered where this pulse is effective in controlling the circuits of the receiving system.

In the receiving system, two sweeps are produced during the time of one recurrence interval of the signals received from the transmitting groups (fig. 13-13). One reset pulse from V37 is required to produce each of the two sweeps. Thus, the output frequency of the reset

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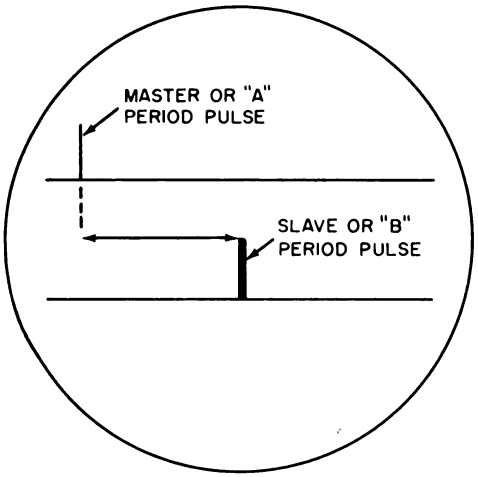

Figure 13-13.—Cathode-ray tube showing A and B period traces.

thyratron, V37 (fig. 13-10) must be twice the receiving system recurrence rate. The purpose of the two sweeps is to allow the indicator circuits to display separate traces for the master and slave pulses, respectively, from the transmitting group.

During operation, the total period of the first trace is considered the A period (fig. 13-3) and the period of the second trace the B period. The master pulse is positioned in the first (A period) trace. The slave pulse, which always occurs at a time greater than 1/2 the transmitting group recurrence rate, is positioned on the second (B period) trace.

The twelfth counter stage, V31 (fig. 13-10) is active in producing the A period and B period traces. The operation of the twelfth counter stage, V31, is the same as that discussed earlier with the use of figure 13-11.

The plate output pulse of the reset thyratron, V37, is coupled by C85 to the junction of the two diodes between the V31 plates, and causes V31 to be triggered into operation. Since two trigger pulses from V37 to V31 (at twice the PRR) are necessary to produce a single square-wave cycle output from V31, the V31 output frequency is equal to the pulse repetition rate of the incoming signals from the transmitting group.

The output of the A section of the twelfth counter, V31A (at the pin 6 plate) is positive during the A period, and negative during the B period. This output is coupled through R145 to the A section grid of 12th amplifier, V38A. V38A functions as a cathode follower to develop its output across R146. The V38A output is fed to delay coincidence stage V45 and to a trace separation amplifier, V62.

The output of the B section of the twelfth counter, V31B, at the pin 4 plate is negative during the A period, and positive during the B period. This output is coupled from the V31B plate through R147 to the B section of the 12th amplifier, V38B. The V38B cathode output across R148 is applied along parallel paths to a gate amplifier V53 (fig. 13-15), and to the B section of an amplitude balance tube, V61 (fig. 13-18). The V38B plate output is coupled by C86 to the A section of the amplitude balance tube, V61 (fig. 13-18). The purpose of the outputs will be considered later in this chapter.

Variable Delay Circuits

The variable delay coincidence stages, V39 through V44 (fig. 13-10) produce a pulse that is continuously variable in delay over the entire B period. The actual delay is read on the time-difference dial on the receiver front panel (fig. 13-1).

As stated earlier, the coincidence action of selected counter outputs of any of the tubes (V20 through V30) with the output of the goniometer circuits (V47 and V48) in the delay coincidence indicator stages (V39 through V45) produces the variable delay. The V39 to V44 output is combined with the output of the twelfth amplifier to produce the variable B pedestal.

It is sufficient at this time to regard the goniometer as a device whose output varies continuously in phase throughout its 100-μs period.

The delay crank (fig. 13-1) controls the position of the lower pedestal (fig. 13-3) within the B period. The delay crank accomplishes this action by operating switch S8 (fig. 13-10) in such a manner that the B pedestal pulse occurs at an interval of time that corresponds to the reading indicated on the time-difference counter.

The delay coincidence indicator, V39, has its A and B sections operating into a common plate load resistor, R149. The grid resistors of V39 (R150 and R151) are returned through a common resistor, R152, to ground. The negative bias at each grid of V39 is determined by the level of the applied signals (from the counter stages) and the amount of the negative

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voltage across R152 as a result of the discharge currents from C87 and C88, respectively. V39 will conduct when either of its grids is driven positive.

S8A through S8D are the sections of S8, which connect the grids of V39 to the plates of V22 in the counter chain when the delay crank is rotated. The common cathode voltage across R152 at V39 grid maintains each section of V39 cutoff when its input signals are momentarily disconnected through the rotating contacts of S8.

V39A receives the grid input from the A plate of the third counter, V22, through S8A and C87. The V39B grid input is coupled from the B plate of the third counter, V22, through S8D and C88. The duration of the third counter output is 100 μs, having positive and negative alternations of 50 μs each. When either section of the tube is conducting, the voltage drop across the common plate load resistor, R149, produces a negative rectangular output pulse, which is 50 μs in width. This pulse is direct-coupled from the V39 plate to the grid of V40.

As the delay crank is turned, S8, which is geared to the crank shaft also turns, and the positive alternation of the input waveform from the third counter is constantly being applied to one of the V39 grids.

The time-difference counter (fig. 13-1) changes at a rate of 100 units per revolution of the goniometer (S8) shaft. One revolution of the goniometer corresponds to 100 μs. A gear reduction of 4 to 1 between the S8 shaft and sections A through D of S8 (fig. 13-10) causes these switch sections to turn 90° for every revolution of the goniometer. Since the goniometer shaft directly drives the mechanical counter, the counter will change 400 units as the output of V39 moves through a 400-μs cycle (one revolution of the S8), and returns to the starting point. Figure 13-14 shows the waveform at the plate of V39 for every 22.5° interval between 0° and 360° points of the S8 drive shaft rotation.

Because the positive grid input is applied to one of the V39 grids at any time throughout the 400-μs interval, one section of V39 will always be conducting. This action causes the V39 negative plate output to be constantly changing position in the 400-μs period.

The switching action of S8A through S8D causes the duration of the V39 output to be alternately 50 and 100 μs, as shown in figure 13-14. This prevents discontinuity at the points on the time-difference counter where the V39 output would change from one interval of the 50-μs pulse to the next.

During a complete revolution of the shaft of S8 (fig. 13-10), the V39 output pulse continuously advances through the entire 400-μs cycle, as stated earlier. The process is repeated as rotation of the delay crank is continued.

Delay coincidence indicator V41 functions the same as V39. The input from the V24A counter to the grid of V41A is applied through sections E and F of S8. The input from V24B to the grid of V41B is fed through section H of S8. The output waveforms of V41 have the same appearance as those from V39, except that the period of the V41 output is alternating 200 and 400 μs as S8 rotates.

A second 4-to-1 gear reduction, which is introduced in sections F through I of S8, causes these switch sections to complete one revolution for every four revolutions of sections A through D of S8, or for every 16 revolutions of the delay crank. Thus, the mechanical counter will change 1600 units as the V41 output moves through a 1600-μs cycle. The cycle is repeated for every 1600 divisions of the mechanical counter, or every 16 revolutions of the goniometer.

As the delay crank shaft is turned, the negative-going output of V41 advances through the entire cycle in the same manner described for the V39 output (fig. 13-14). The pulse occupies each 200-μs interval, in turn, throughout the 1600-μs period.

V42 and V44 (fig. 13-10) operate in the same manner as V39 and V41, just discussed. The input to the V42 grids from the seventh counter stage, V26, is controlled by sections I through L of S8. These switch sections are also driven by the S8 shaft through a third 4-to-1 gear reduction. The output pulse width of V42 is 800 μs, but alternately doubles to 1600 μs as the switch, S8, rotates. V42 couples its output to V43B.

Due to the third 4-to-1 gear reduction, the V42 cycle repeats itself once for every 6400 scale divisions on the counter, or once for every 64 rotations of the delay crank. As the crank rotates, the V42 output advances through the 6400-μs period, and the position (in time) of the V42 output is also advanced.

The fourth 4-to-1 gear reduction takes place in sections M, N, P and Q of S8. The total reduction of these sections of S8 is 25,600 to 1. Thus, one revolution of sections M, N, P and Q of S8

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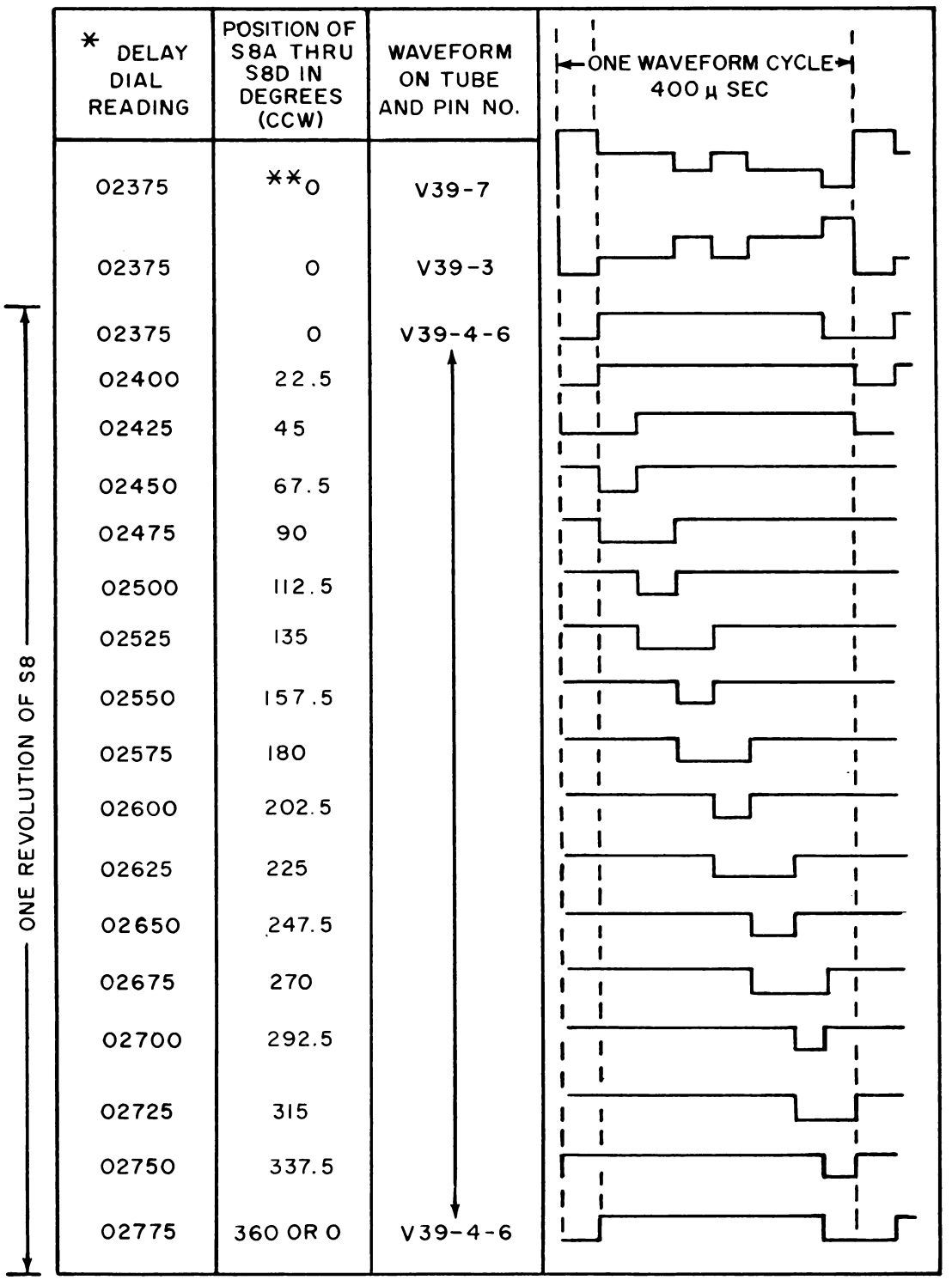

* REPEATS IN 400 μ SEC.

** SWITCH POSITION AS SHOWN ON SCHEMATIC DIAGRAM

Figure 13-14.—Gating waveforms from delay coincidence indicator V39.

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corresponds to a 25,600-division change on the mechanical counter.

Sections M, N, P and Q of S8 control the input of the grids of the A and B sections at V44 from the ninth counter stage, V28. The positive duration of the V28 (ninth counter) output is 3200 μs. The input to the grid of V44 alternately changes from 3200 to 6400 μs as S8 rotates. Due to the fourth 4-to-1 gear reduction, the V44 grid input cycle repeats itself after a 25,600-μs period. The V44 plate output is applied to the grid of V43A.

The delay coincidence stages, V40, V43, and V45, have their plates connected through common load resistors R158 to the +160-volt supply. The cathodes of V40, V43, and V45 are returned to the +36-volt tap on the divider comprising R153, R154, and R155.

Plate current flows in V40 unless both A and B sections of the tubes are cut off at the same time. This condition only occurs when V39 and V41 are conducting simultaneously. Likewise, V43 cuts off only when its grid input from V42 and V44 are negative going (V42 and V44 conducting).

The pulse period of the voltages simultaneously applied to the grids of V40 and V43 varies as a result of the combined inputs from V39, V41, V42, and V44, as the delay crank is turned through the entire delay period from 0 to 25,600 μs. Because V40, V43, and V45 are parallel connected and conducting through the same load (R158), the plate voltage of either of these tubes does not rise appreciably, unless all of the tubes (V40, V43, and V45) cut off. V40 and V43 are cut off by the negative inputs from V39, V41, V42, and V44, as discussed.

The V45 grid receives the output from the cathode of the A section of the twelfth amplifier, V38. It should be recalled that the output of the A section of V38 was positive during the A period. Thus, during the A period a positive pulse is applied to the V45 grid through R157 and C89. This potential keeps V45 conducting throughout the A period, and no rise in plate voltage from either of the tubes (V40 or V43) can be obtained.

During the B period, the output at the V38A cathode is negative going. When this potential is present at the grid of V45 (via R157 and C89), V45 is cut off, and in the presence of the V40 and V43 cutoff, the plate voltage of all the parallel tubes (V40, V43, and V45) rises to apply a positive-going pulse to a delay thyratron tube, V49 (fig. 13-15). Since the plate output of V39 is negative for either 50 or 100 μs, as discussed, V40A will begin conduction after this period, thereby terminating the delay coincidence indicator output pulse.

S8R, S8T, and S6J (fig. 13-10) are used as an electrical stop for the variable delay coincidence system. When the PRR switch, S6, is in any of the L recurrence interval positions (see table 13-1), a -108-volt source is connected through S6J (contacts 14 and 23) to contact 5 of S8T. When the delay crank is turned slightly beyond the delay corresponding to one-half of the longest recurrence interval, the -108 volts is applied to the grids of V39 through contacts 5 and 14 of S8T in the position shown. This negative voltage is sufficient to hold V39 beyond cutoff for any signals arriving at the V39 grids, and thereby disables the variable coincidence chain. With V39 in the nonconducting state, V40 will draw current continuously, and no rise in plate voltage (output pulse) of the coincidence indicator output tubes (V40, V43, and V45) can occur.

S6J and S8R operate together in a similar manner to hold V41 beyond cutoff when the PRR switch, S6, is in one of the H positions (table 13-1). The stop is not used in the S positions of the PRR, S6. The stop is also effective for a slight period of time after 98,375 on the counter dial before the zero point is reached on the H and L rates.

Goniometer and Clipper Circuits

The output of the third counter stage, V22 (fig. 13-10) is coupled by R169 and C97 to the control grid of a shaper stage, V50A (fig. 13-15). Without regard to the square-wave input, C97 initially charges to the plate potential of V22A (fig. 13-10). The grid of V50A (fig. 13-15) would begin to draw a current because of the connection of the V50A grid through R168 to the +160-volt supply. With the square wave applied, C97 charges toward the V22A (fig. 13-10) plate potential through the grid-cathode resistance of V50A (fig. 13-15), which forms a short, time-constant charge path. The discharge path of C97 is through R168 (3.9 megohms), which makes up a long, time-constant discharge path, and a negative grid-to-cathode (bias) voltage is established at the V50A grid.

V50A amplifies the square-wave input and applies its output through R165 and R163 to a 10-kc filter comprising R162, L5, C92, and C91. The filter output is a 10-kc sine wave, which is

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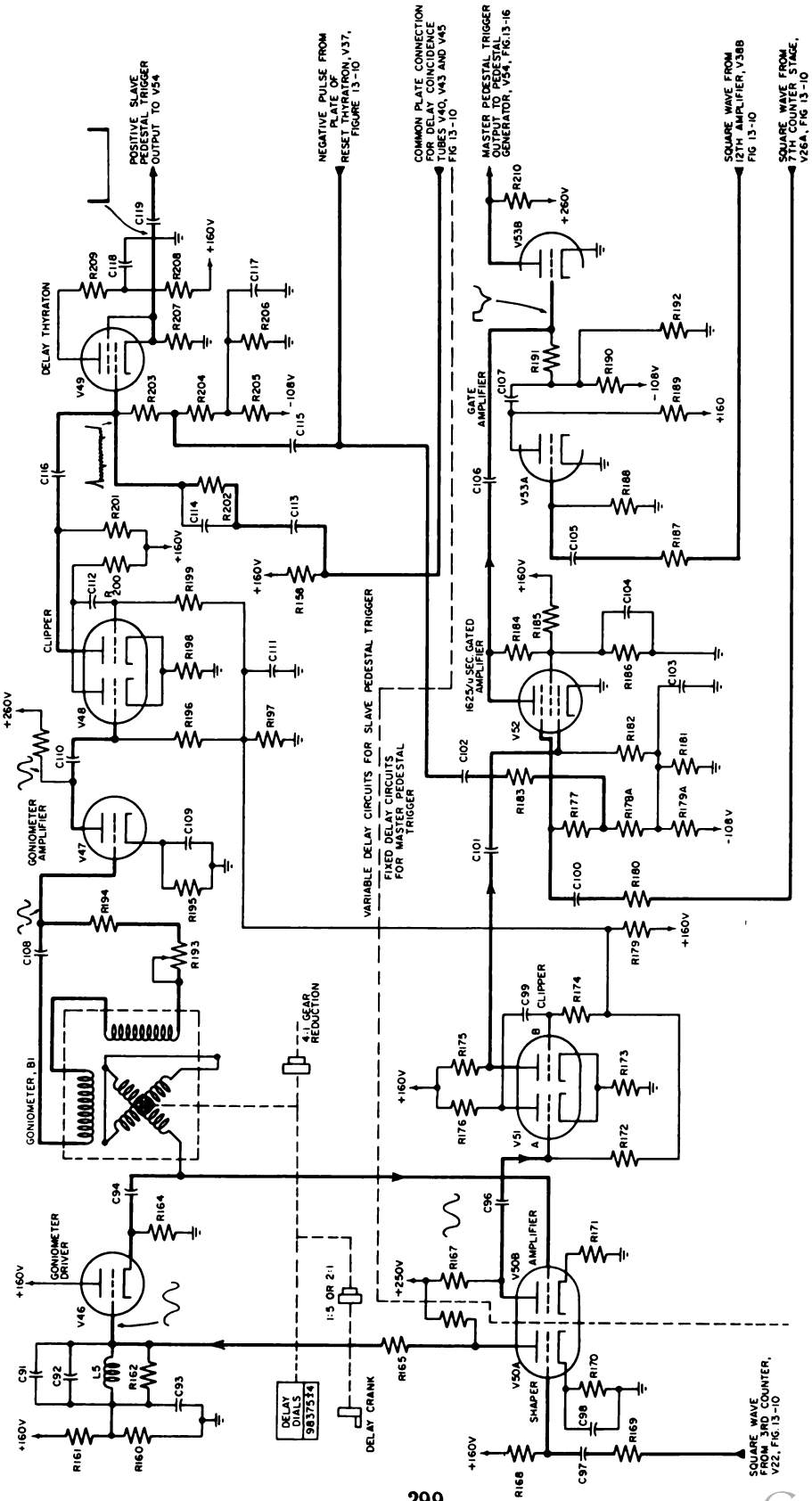

Figure 13-15.—Goniometer, clipper, and fixed delay circuits.

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applied to the grid of the goniometer driver, V46.

Goniometer driver V46 is a cathode follower with a positive grid bias. The positive grid potential is derived from the L5-R163 junction, which forms a portion of a voltage divider comprising R170, V50A, R165, R163, L5, and R161 connected from the +160-volt supply to ground. The cathode potential of V46, as a result of the V46 tube conduction through R164, is about +42 volts. These potentials hold V46 at the proper operating point.

The output of V46 across R164 is a 10-kc sine wave, which is coupled by C94 to the goniometer, B1. Thus, the period of the goniometer input is 100 μs, the same as that of the widest pedestal pulse from the delay coincidence indicator, V45 (fig. 13-10).

The goniometer driver (V46) output (fig. 13-15) is also coupled to the grid of the 10-kc amplifier, V50B. The purpose of this input is considered later.

The goniometer (B1) output has a constant peak-to-peak amplitude per cycle, but is continuously variable in phase with respect to its input. The goniometer is geared 1:1 to the counter, and both the goniometer and counter are geared 4-to-1 to the shaft that drives S8 (fig. 13-10). One revolution of B1 (fig. 13-15) moves the B1 output through one complete cycle (100 μs), and the counter through 100 scale divisions.

C108, R194, and R193 provide the proper impedance for terminating the goniometer (B1) output. R193 is the goniometer balance potentiometer, which is adjusted to balance the quadrature relationship of the goniometer output.

The goniometer (B1) output from the C108-R-194 junction is applied to the grid of goniometer amplifier V47. The amplified output of V47 is applied through C110 to the grid of a clipper stage, V48.

The clipper stage, V48, functions similarly to a cathode-coupled multivibrator. The grids of V48A and B are tied through R196 and R199 respectively, to a +17-volt tap on a voltage divider comprising R197 and R179. The cathode potential across common cathode resistor R198 is +26 volts. Positive portions of the input sine waves from V47 cause conduction of V48A. Negative portions of the V48A input cause cutoff of this tube section.

The B section of V48 is coupled to V48A, through the common cathode resistance, R198, and by coupling capacitor C112. V48B is cut off for negative inputs and conducts for positive inputs. The clipper (V48) output is not a square wave because of the short time constant of C116, R203, R204, and R205. The input to the delay thyratron, V49, from V48 is therefore a differentiated pulse. Only the sharp, positive differentiated pulse is effective at the V49 grid, as will be seen later.

The delay thyratron, V49, receives two other inputs simultaneously with the application of the clipper (V48) input. The negative pulse from the plate of the reset thyratron, V37 (fig. 13-10) is applied through C115 and R203 (fig. 13-15) to the V49 grid. The positive output of the delay coincidence tubes, V40, V43, and V45 (fig. 13-10) is fed through C113 and C114 (fig. 13-15) to the grid of V49. C113 and C114 in parallel with R202, maintain the shape of the pedestal pulse, and also isolates the delay coincidence indicator stages (fig. 13-10) from the clipper stage output (fig. 13-15).

The position (in time) of the positive output of the clipper stage, V48, to the delay thyratron (V49) grid varies continuously as the goniometer shaft is turned. The positive trigger, which is applied to the V49 grid from V48 in the presence of the positive delay coincidence input to V49 causes the delay thyratron to produce a positive-going slave pedestal trigger pulse. Because the delay coincidence input pulse from V45 (fig. 13-10) is only available during the B period, and because the input pulse from the goniometer (fig. 13-16) is continuously moving through a 100-μs period as the goniometer shaft is turned, the time of the conduction of V49 can vary throughout the 100-μs interval.

The exponential slope of the input from the reset thyratron, V37, as discussed (fig. 13-10) is sufficiently negative to keep the delay thyratron, V49 (fig. 13-15) from firing during the short interval after reset of the counter stages (V20 through V31 of figure 13-10) when many transients are present. These transients could be coupled through the delay coincidence and goniometer stages with sufficient amplitude to false-trigger the delay thyratron, V49 (fig. 13-15) even in the absence of the normal delay coincidence indicator chain output. The delay period of the V49 firing introduced as a result of the negative reset thyratron pulse is small. Because zero delay for the receiving system occurs an A delay period after the reset (fig. 13-3), the accuracy of the delay reading is not affected.

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Note that the delay pedestal from the parallel plates of the delay coincidence stages, V40, V43, and V45 (fig. 13-10) is not the slave pedestal pulse and is not the pulse displayed on the indicator tube (fig. 13-3). Rather, the positive output pulse of these stages is used at the grid of V49 (fig. 13-15) to produce the slave pedestal trigger at the V49 cathode.

The V49 output across R207 is coupled by C119 to the slave pedestal generator circuits (fig. 13-16). The pedestal generator also receives a master pedestal trigger pulse from the fixed delay circuits (fig. 13-15). The method by which the master pedestal trigger is generated will be considered before discussing the generation of the master and slave pedestals.

The pedestal generator stages, V54 and V55 (fig. 13-16) produce the pedestals as they are seen on the indicator screen (fig. 13-3).

Fixed Delay Circuits

The trigger pulse used to initiate the master pedestal is produced in the fixed delay circuits (fig. 13-15) by combining a pulse obtained from the leading edge of the sine-wave output of the 10-kc amplifier (after passing through clipper stage, V51) with square waves from the seventh counter stage, V26, and the twelfth amplifier, V38 (fig. 13-10). These inputs produce the master pedestal trigger pulse an A delay period after reset in the A trace period (fig. 13-3). Thus, the time of the master pedestal trigger corresponds to the leading edge of the master pulse, which begins 1625 μs after the beginning of the A period.

In operation, the time-difference counter is adjusted to give a zero reading 1625 μs after the beginning of the A period. Thus, any change in the time-difference counter reading from zero in the B period trace (when the difference is being determined) represents the actual difference between the master and slave pulses less one-half the recurrence interval.

The 10-kc amplifier, V50B (fig. 13-15) receives the goniometer driver (V46) output through C94 and couples an amplified 10-kc sine-wave output through C96 to the grid of clipper stage V51A. The clipping action in V51 is produced in the same manner as that described in the V48 clipper in the variable delay circuits. The 10-kc square-wave output of V51B is differentiated by C101, R182, and R181, and applied to the control grid of the 1625-μs gated amplifier, V52.

The 1625-μs gated amplifier, V52, also receives a 1600-μs square wave from the seventh counter stage, V26A (fig. 13-10). This input is applied through R180 and C100 (fig. 13-15) to the V52 suppressor grid. V52 is a coincidence stage, which requires the simultaneous application of both input pulses (from V51 and V26A) before conduction of V52 will occur.

The time constant of the coupling network to the V52 suppressor grid is long enough to hold the suppressor at approximately zero volts for a 25-μs interval before the clipper (V51) output is applied to the control grid of V52. When the V51 output arrives at the V52 control grid, a negative output pulse is produced at the V52 plate.

The negative input pulse, coupled through C102, R183, and R177 to the suppressor grid, from the reset thyratron, V37 (fig. 13-10) prevents firing of the 1625-μs gated amplifier, V52 (fig. 13-15) due to transient voltages present immediately after the reset period. The 1625-μs gated amplifier, V52 (fig. 13-15) produces an output pulse on the A delay period after each reset pulse and at 1600-μs intervals following the first output.

The gated amplifier, V53B, is normally held cut off by a -20-volt potential obtained from the junction of R190 and R192. The output pulse of the 1625-μs gated amplifier, V52, is applied to the control grid of V53B through C106. This input pulse alone is not sufficient to cause the conduction of V53B.

The control grid of V53A receives and amplifies the square-wave output of the twelfth amplifier, V38B (fig. 13-10), which is negative during the A period and positive during the B period. Since in V53A (fig. 13-15) the input pulse is inverted at the V53A plate, the V53A output is positive going during the A period. The time constant of C107 and R192 (in the V53A output coupling network) maintains the V53B grid above cutoff just long enough for the first input pulse from the 1625-μs gated amplifiers to the V53B grid (in the A period) to be amplified in this stage. The V53B output represents the master pedestal trigger output pulse to the pedestal generator circuits (fig. 13-16).

Pedestal Circuits

The trigger pulses produced in the fixed and variable delay circuits, as discussed (fig. 13-15) are used in the pedestal generator stages, V54 and V55 (fig. 13-16) to initiate the pedestal pulses

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seen on the indicator screen (fig. 13-3). The master and slave trigger pulses are applied (during separate periods) to the grid of pedestal generator stage, V54 (fig. 13-16). The circuit of V54 is that of a cathode-coupled multivibrator and operates similarly to the A and B multivibrators (fig. 13-7) discussed earlier.

The pedestal generator stage, V54B (fig. 13-16) is normally conducting because its grid is connected through R214 to the +260-volt supply. The pedestal trigger pulses (either master or slave from V49 or V53B, respectively, of figure 13-15) are applied to the V54A grid (fig-13-16), causing this tube section to conduct. The resulting positive-going output pulse of V54B is coupled by C123 to the pedestal amplifier (V55) control grid. The positive output at the V55 cathode is coupled to the AFC circuits (fig. 13-7) to initiate the action of the A multivibrator, V8.

The sweep function switch, S7 (fig. 13-16) controls the manner in which the pedestal pulses will be used and the nature of the pattern shown on the indicator screen. The modes of the switch, referred to as sweep function 1, sweep function 2, and sweep function 3, describe the manner of the indicator display and the use of the pedestal pulses rather than the position of the switch contacts (fig. 13-16). The cathode-ray display for each of the S7 modes is shown in figure 13-17. The method by which the three different patterns are produced is discussed presently.

Trace Separation

The output of the twelfth counter stage, V38A (fig. 13-10) is fed into the trace separation amplifier, V62 (fig. 13-16). The purpose of the trace separation amplifier, V62, is to produce a potential (at the V60 vertical deflection plates) which causes the electron beam to be presented in the upper section of the indicator (V60) tube screen during the A period, and towards the lower section of the V60 screen during the B period.

The output of V38A (fig. 13-10), as stated earlier, is positive during the A period and negative during the B period. The V38A output is coupled to the grid of V62 (fig. 13-16) through C136 and R257.

When the equipment is initially set into operation, positive portions of the V38A output cause the V62 grid to draw current, thereby charging C136 negative toward the V62 grid.

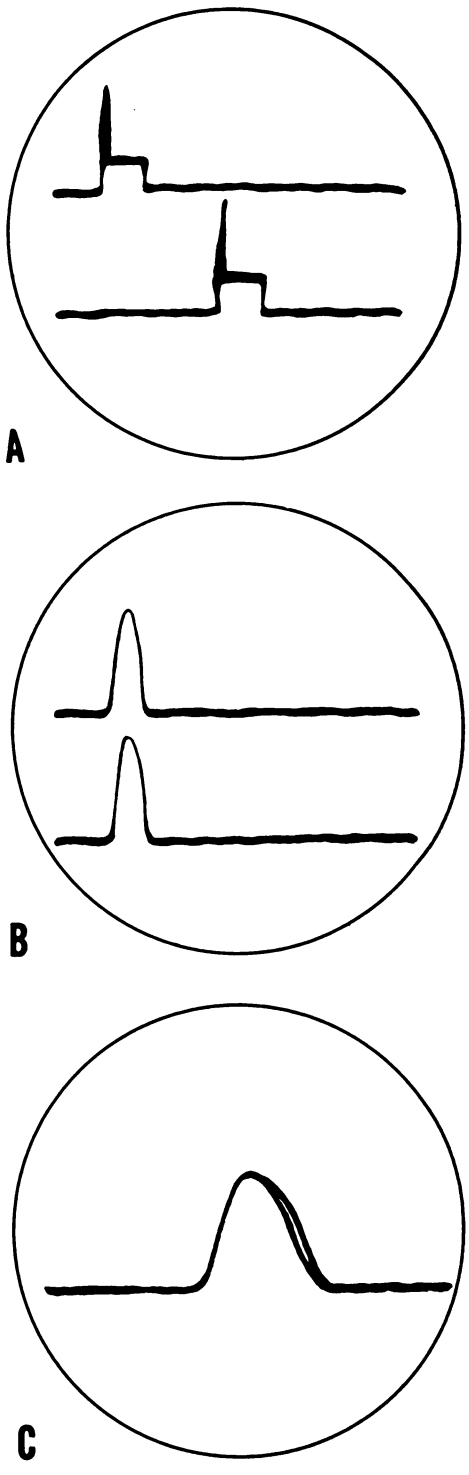

Figure 13-17.—Cathode ray tube presentations for sweep functions 1, 2, and 3.

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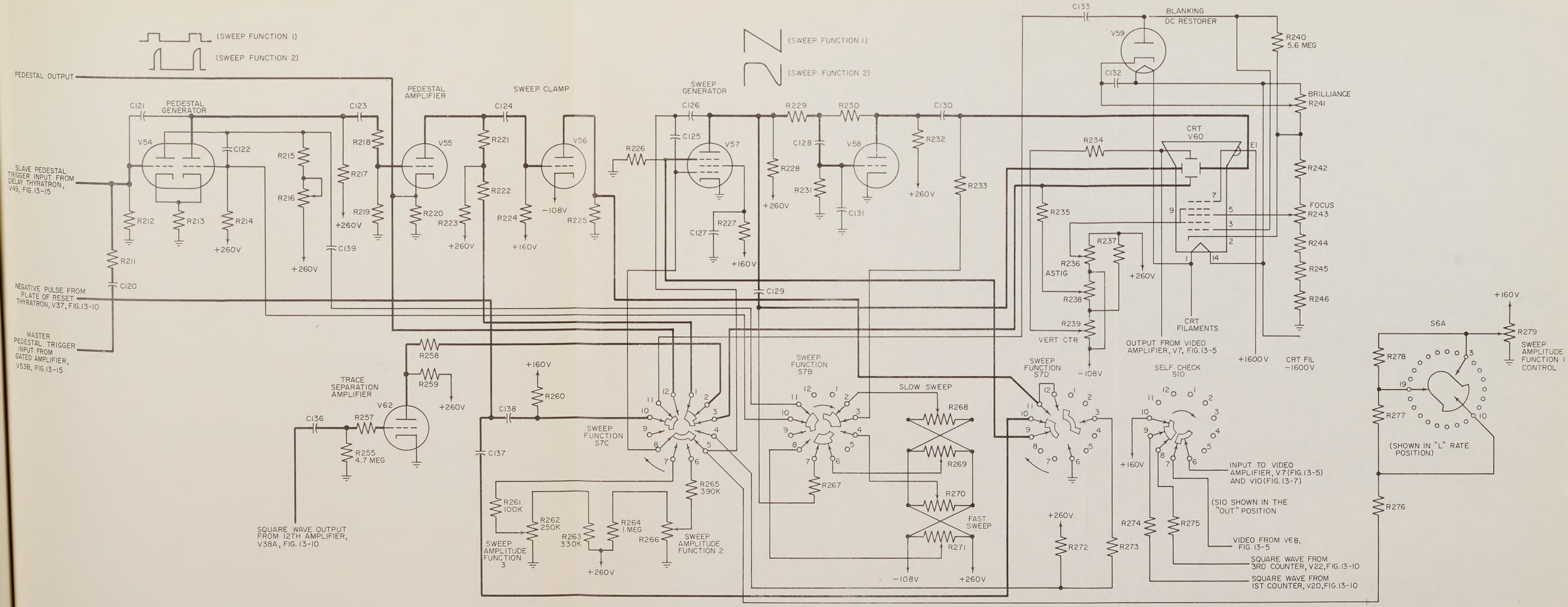
(SWEEP FUNCTION 1)
(SWEEP FUNCTION 2)
PEDESTAL OUTPUT
SLAVE PEDESTAL TRIGGER INPUT FROM DELAY THYRATRON, V49, FIG. 13-15
NEGATIVE PULSE FROM PLATE OF RESET THYRATRON, V37, FIG. 13-10
MASTER PEDESTAL TRIGGER INPUT FROM GATED AMPLIFIER, V53B, FIG. 13-15
PEDESTAL GENERATOR
PEDESTAL AMPLIFIER
SWEEP CLAMP
SWEEP GENERATOR
BLANKING DC RESTORER
CRT V60
BRILLIANCE R241
FOCUS R243
ASTIG
CRT FILAMENTS
OUTPUT FROM VIDEO AMPLIFIER, V7, FIG. 13-5
VERT CTR
+1600V
CRT FIL −1600V
TRACE SEPARATION AMPLIFIER
SWEEP FUNCTION S7B
SWEEP FUNCTION S7C
SWEEP FUNCTION S7D
SLOW SWEEP
FAST SWEEP
SELF CHECK S10
S6A
SWEEP AMPLITUDE FUNCTION 1 CONTROL
(SHOWN IN "L" RATE POSITION)
SWEEP AMPLITUDE FUNCTION 2
SWEEP AMPLITUDE FUNCTION 3
SQUARE WAVE OUTPUT FROM 12TH AMPLIFIER, V38A, FIG. 13-10
INPUT TO VIDEO AMPLIFIER, V7 (FIG. 13-5) AND V10 (FIG. 13-7)
(S10 SHOWN IN THE "OUT" POSITION
VIDEO FROM V6B, FIG. 13-5
SQUARE WAVE FROM 3RD COUNTER, V22, FIG. 13-10
SQUARE WAVE FROM 1ST COUNTER, V20, FIG. 13-10

Figure 13-16.—Pedestal, trace separation, blanking, and self-check circuits.

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The accumulated charge of C136 cannot be lost entirely through R255 (4.7 megohms) during negative portions of the V38A input, and approximately -13 volts is maintained between the grid and cathode of V62. Thus, after the bias is established, V62 amplifies and inverts the input square wave from V38A (fig. 13-10).

In sweep function 1 (as shown) the V62 output (fig. 13-16), which is negative going during the A period and positive during the B period, is coupled through R258 and contacts 2 and 3 of S7C to the lower deflection plate of cathode-ray indicator tube, V60. Since this output is applied to the lower vertical deflection plate of V60, the V60 electron beam is repelled by the negative-going V62 output during the A periods, and the A period trace is therefore presented on the upper section of the V60 screen. During the B period, the V62 output is positive going, and the B period trace will be attracted downward toward the now more positive lower deflection plate.

Note that during sweep function 1, the pedestal amplifier (V55) plate output (which is also negative going as discussed) is applied through R221 and R222, and through the 1-3 contacts of S7C to be superimposed on the trace separation output of V62. Because the pedestal pulses from the V55 plate are also negative-going voltages, the potential on the lower deflection plate of V60 is made more negative for the duration of the pedestal pulses in the A and B period traces, and the pedestal pulses are presented to the V60 observer (fig. 13-17, A).

Sweep Circuits

In sweep function 1, S7 (fig. 13-16) is in the position shown, and the negative-going plate output of the reset thyratron, V37 (fig. 13-10) is coupled through C137 (fig. 13-16) through contacts 9 and 10 of S7D to the suppressor grid of sweep generator stage V57. The reset thyratron (V37) plate output (fig. 13-10) goes negative at reset, and rises exponentially (due to the change of C84) when V37 again cuts off toward the B voltage value (+160 volt). The negative drop in potential that is applied to the V57 suppressor grid, as discussed (fig. 13-16) holds the suppressor sufficiently negative to cut off the V57 plate current. Thus, C126 at the plate of V57 charges toward the B supply potential through R279 and R278, through contacts 19 and 10 of S6A (in the position shown) and through R276.

Since the time constant of the charge circuit of C126 is short, C126 will become completely charged by the time that the V57 suppressor input rises sufficiently to allow V57 to again conduct. The suppressor potential goes slightly positive at the end of the negative reset pulse, as a result of the plate potential rise of V37 (fig. 13-10) when V37 is cut off. The path for the V37 plate voltage rise is through R226 (fig. 13-16) through the 9 and 10 contacts of S7C and through C137 to the V37 plate (fig. 13-10).

When the V57 suppressor grid (fig. 13-16) is driven positive, plate current will again flow. The resulting drop in the V57 plate voltage and rise of plate current cannot occur at a faster rate than that required for C126 to lose its accumulated charge. The C126 discharge path is through R276, through contacts 10 and 19 of S6A (in the position shown), through R278, and through the arm-to-the-B-supply resistance of R279, returning to C126 through the d-c conducting resistance of V57. The negative potential thus developed at the V57 grid from R276 to ground is applied to the V57 control grid, and opposes the rise in the V57 plate current. In addition, the connection of C126 and C125 between the grid and plate of V57 makes the input capacity of the stage equal to the value of the grid plate capacity of V57 multiplied by the stage gain. The high grid-to-ground capacity of V57 permits the grid voltage to change only at a very slow rate. Consequently, the effective time constant for the discharge of C126 is sufficiently large to hold the negative-going V57 plate output linear over the duration of the positive suppressor input.

The resulting V57 plate output is a linear-rising sawtooth waveform. The sawtooth sweep duration during sweep function 1 is approximately 1,500 to 25,000 μs, depending upon the PRR switch, S6 (fig. 13-10).

The sweep amplitude function 1 control, R279 (fig. 13-16) adjusts the magnitude of the positive-voltage applied to the V57 control grid, thereby locating the operating point of the tube and determining the amplitude (and width) of the V57 plate sawtooth output. For L rates (see table 13-1) R277 (fig. 13-16) is shorted by contacts 10 and 19 of S6A (as shown). For S rates (fewer cycles per second than for L rates) both resistors, R277 and R278, are in the C126 discharge path, and the discharge time is increased. Thus, fewer sawtooth voltage waveforms per second will be produced by V57. For H rates (more cycles per second, see table

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13-1) contacts 10 and 3 short R277 and R278, and the sawtooth output frequency is increased.

In sweep function 2 each of the switch sections (S7B, and S7D) is rotated 1 contact to the left from the position shown. For sweep function 3, the switch sections are rotated 2 contacts to the left.

During sweep function 2 and sweep function 3, the pedestal pulses are not displayed on the indicator screen (fig. 13-17). Instead, the pedestals are coupled through C124 to the grid of the sweep clamp stage, V56. The negative-going grid input is amplified in this stage.

Because of the larger negative voltages at the V56 cathode (-108 volts) V56 will conduct and amplify strong negative signals. The V56 grid is negative to the cathode due to the discharge of C124 through R224.

During sweep function 2 the V56 plate pedestal output is coupled through contacts 11 and 9 of S7D to the suppressor grid of the sweep generator, V57. The sweep amplitude function 2 control, R266, is selected by contacts 8 and 6 of S7C, and is used to adjust the sawtooth amplitude for the sweep function 2 mode of operation. The sweep duration for sweep function 2 is approximately 1250 μs.

For sweep function 3 operation, the sweep clamp (V56) plate pedestal output is coupled through contacts 12 and 9 of S7D to the sweep generator (V57) suppressor grid. The sweep amplitude function 3 control, R262, which is used to adjust the V57 plate sawtooth output amplitude, is selected by contacts 8 and 7 of S7D. The duration of the sweep for sweep function 3 operation is about 160 μs.

As stated earlier, R279 functions to adjust the sweep amplitude (or sweep duration) for sweep function 1. The sweep duration is different for each position of the sweep function switch, S7. To obtain adequate sweep amplitude range for each of the sweep functions (1, 2, and 3) the time constant of the C126 discharge must be changed.

For sweep function 2 (S7 one position to the left) C125 is placed in series with C126, and the discharge path of C126 is through contacts 8 and 6 of S7C, through R265, through the upper section of R266 (from the R266 arm) through R264, and the B supply to ground, returning to C126 through the d-c conducting resistance of V57. The resistors in the discharge path during sweep function 3 selected by the 8 and 7 contacts of S7C included R261, R262 (upper section), and R263. Thus, R266 and R262 adjust the sweep amplitude for sweep function 2 and sweep function 3 modes of operation, respectively.

The output of the sweep generator, V57, is coupled through C129 to the left horizontal deflection plate of the cathode-ray tube, V60. The V57 output is also coupled through R229 and C128 to the control grid of the sweep inverter, V58.

The sweep inverter (V58) grid is grounded through C131. This capacitor bypasses the grid resistor, R231, to reduce differentiation of the input, which would otherwise occur across R231, thereby altering the shape (rise time) of the sawtooth input. The V58 input sawtooth is inverted in the stage, and coupled through C130 to the right horizontal deflection plate. Inverse feedback through R230 from plate to grid of V58 is used to hold the amplitude of the V58 output equal to the output of V57, thus supplying the same magnitude voltage to the left and right horizontal deflection plate.

Blanking Circuits

The blanking d-c restorer, V59 (fig. 13-16) prevents the appearance of the sweep voltage retrace on the V60 screen. V59 is connected across the potential difference existing between the R241 and R242 junction (at the V59 plate) and the arm of R241 (at the V59 cathode). These resistors form a portion of a voltage divider from a -1600-volt potential to ground. The plate of V59 is positive to the cathode, and the tube conducts. C133, at the plate of V59, charges from the plate potential (approximately -1600 volts) through contacts 11 and 10 of S7C, and through R260 toward the +160-volt supply.

It will be recalled that the negative pulses arriving from the plate of the reset thyratron, V37 (fig. 13-10) terminate the sawtooth waveform at the V57 plate (fig. 13-16). At the same time, the reset thyratron negative plate output is applied through C138 and contacts 10 and 11 of S7C to C133. Thus, during the period of the negative trigger, C133 discharges through R240, developing a large negative potential at the control grid (pin 3) of the cathode-ray tube, V60. This action blanks the indicator tube during the periods of the sawtooth waveform flyback.

Cathode-Ray Tube Circuits

The accelerating potentials required for the operation of the cathode-ray tube, V60 (fig. 13-16) are obtained from separate positive and negative supplies (not shown). Both supplies

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evelop a 1600-volt output. The positive supply connected to the third anode, E1, of V60.

A voltage divider comprising R241 through 246 is connected between the -1600-volt source nd ground. Potentials obtained from this di-ider provide the voltages required at the athode (pin 2) control grid (pin 3) and second node (pin 5) of V60. The control grid potential determined by the setting of the arm of the rilliance control, R241, and the resulting con-uction of V59. The second anode is connected the focus control, R243, and the cathode to ne junction of R241 and R242.

Since the cathode of V60 is highly negative ith respect to ground, the heater (between pins and 14 of V60) is returned to the -1600-volt upply to keep a low potential difference between ne heater and cathode of V60. The heater of V59 s also returned to the -1600-volt supply for ne same reason.

The voltage at the first anode (pin 9) of 60 is controlled by the astigmatism adjustment, 236. This control is connected between the 260- and -108-volt supplies.

R238 and R239 are the vertical centering ontrols. R237 in series with the vertical cen-er potentiometers, R238 and R239, limit the oltage range of the controls. The connection of he vertical controls between the -108- and 260-volt supplies (in series with R237) pro-ided push-pull vertical centering. The astig-natism control, R236 functions in conjunction ith the focus control, R243, to determine the efinition of the presentation.

The d-c potentials at the vertical deflection lates of V60 are determined by the settings of 238 and R239, as discussed. The d-c po-entials on the horizontal deflection plates de-end on the position of the sweep function switch ection, S7B. For sweep function 1 (S7B as hown) the settings of R268 and R269 (slow and ast sweep controls, respectively) determine he potential on their respective horizontal late. Each of these resistors is connected be-ween the -108- to +260-volt supplies. For weep functions 2 and 3, the horizontal de-lection plate potentials are obtained from R270 nd R271, respectively. Since the sweep period luring sweep function 1 is considerably longer 15,000 to 25,000 μs) than the sweep used for weep functions 2 and 3 (1250 and 160 μs, re-pectively), the use of two sweep controls (fast nd slow) eliminate timely readjustment of the orizontal plate potentials when switching from ne sweep function to another.

Self-Check Circuit

The self-check switch, S10 (fig. 13-16) permits markers to be produced on the indicator tube so that the sweep length and the alignment of the goniometer and time-difference counter can be checked. When S10 is in the OUT position (as shown) the video output from V6B (fig. 13-5) is connected through contacts 7 and 6 of S10 (fig. 13-16) to the video amplifier, V7 (fig. 13-5), and to the video amplifier, V10 (fig. 13-7).

When S10 (not shown on receiver front panel of figure 13-1) is rotated one position to the left, 100-μs marker pulses are coupled from the output of the third counter stage, V22 (fig. 13-10) through contacts 8 and 6 of S10 to the video amplifier, V7 (fig. 13-5). This input is subsequently applied to the upper vertical deflection plate of the cathode-ray tube.

When S10 is rotated two positions to the left 25-μs markers are obtained from the first counter stage, V20 (fig. 13-10). These pulses are similarly displayed on the cathode-ray tube.

Left-Right Switch

In operation, received signals at the selected pulse repetition rate are locked in on the indicator screen by use of either the manual or automatic frequency control circuits, as discussed earlier. The L-R (left-right) switch, S9 (fig. 13-10) enables the operator of the receiving set to position the signals at any desired location on the trace without having to disturb any of the other controls.

When S9 is in its normal (OFF) position as shown, the switch is not effective in controlling the position of the indicator pattern. As discussed earlier, the reset pulse from the cathode of V37 is applied through the 1 and 3 contacts of S9A and through R106 to the B grid of the first counter stage, V20. The V37 cathode reset pulse is also fed from contact 3 of S9A through contacts 4 and 3 of S7A to the B grid of the fourth counter, V23. Since the B grid of all the other counter stages are permanently controlled by the positive reset pulse, all of the counter stages are then reset on the same side, and the counters, V20 through V31, produce the proper output count selected by the PRR switch. This represents the normal action of the reset and counter stages.

When the L-R switch (section S9A) is operated to the RIGHT position in sweep function 1,

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the positive reset pulse from the V37 cathode is coupled through the 1 and 2 contacts of S9A to the A grid of the first counter stage, V20. The reset pulse is also coupled from contact 2 of S9A, through contacts 10 and 7 of S7A, to the A grid of the fourth counter stage, V23. Thus, the first and fourth counters (V20 and V23, respectively) are reset to the side opposite from normal.

Each counter stage normally produces one output pulse for each two input triggers. However, when the first and fourth counters are reset at the opposite side, the first input trigger pulse from the preceding stage (or from the clipper stage, V19 in figure 13-9) after the reset pulse will cause the first and fourth counters to produce an output. After the first output pulse of V20 and V23 is produced as a result of the action discussed, these stages will produce an output on each odd pulse input.

Because the first and fourth counters are driven by trigger pulses with repetition rates of 12.5 and 100 μs, respectively, the count selected by the PRR switch, S6, will be reached 112.6 μs earlier than normal, and the receiving set repetition rate is higher than that of the received signals. The signals will appear at a later time on successive traces, and drift across the screen to the right.

The sweep time for sweep functions 2 and 3 is much less than in sweep function 1, and a slower drift is necessary to permit the operator to control the received signals. For sweep functions 2 and 3, contact 4 of S7A is open. Likewise, the path for the reset pulse input to the fourth counter, V23, is also open. As a result, only the first counter, V22, is reset to the opposite side, and the selected count is reached only 12.5 μs earlier than normal. This produces a much slower drift of the received signal across the indicator screen.

When the L-R switch, S9B (fig. 13-9) is operated to the LEFT position in sweep function 1 (contacts 2 and 3 connected), the negative plate output of the reset thyratron, V37 (fig. 13-10) is coupled through C141 (fig. 13-9), and through contacts 2 and 3 of S9B to the control grid of the ringer tube, V18. The input pulse is differentiated by the network comprising C141, R281, R280, and C140. The negative input holds the ringer tube, V18, beyond cutoff for a period of nine driving pulses (112.5 μs). The receiving set pulse repetition rate is therefore made lower than that of the received signals. This action causes the received signals to appear at an earlier time on successive sweep traces, and drift across the screen to the left.

C140 normally charges negative to ground during the period of the negative input pulse. The C140 discharge through R280 and R281 (after the trailing edge of the negative input) maintains the negative cutoff potential at the grid of V18 for the required period.

For sweep functions 2 and 3, the cutoff time of V18 is reduced as necessary to produce a slower drift of the received signals. During sweep functions 2 and 3, C140 is shorted to ground by contacts 11 and 8, and 12 and 9, respectively, of S7A (fig. 13-10). The cutoff period of V18 (fig. 13-9) is reduced to that equal to 1 driving pulse (12.5 μs). Thus, the drift toward the left end of the cathode-ray tube is made slower.

Amplitude Balance

The amplitude balance stage, V61 (fig. 13-18) provides a means of equalizing the amplitude of the received master and slave pulses during the A and B periods, respectively. The cathodes of V61 are tied together and returned to the -108-volt supply through R254. The plates of V61 are connected through separate sections of R249, each section of which is in series with R256 to ground.

Both sections of V61 are normally conducting, and the plate potential of the A and B sections of V61 is negative to ground. The negative voltage at the terminal 2 arm of R249 is applied to the control grid of the r-f and i-f stages, V1, V3, and V4 (fig. 13-5).

The twelfth amplifier (V38B) cathode output (fig. 13-10) is applied to the A grid of V61 (fig. 13-18) and is negative during the A period, and positive during the B period. The V38B plate output (fig. 13-10) is fed to the grid of V61B, and is positive during the A period and negative during the B period. The width of the twelfth amplifier (V38) output is equal to the width of the recurrence interval of the received signals, with the positive and negative alternations equal to one-half the total recurrence period.

The potential at terminal 2 arm of R249 for the separate halves of the twelfth amplifier input depends upon the setting of the terminal 2 arm. If the terminal 2 arm is moved to the right or left, the negative potential fed to the r-f and i-f stages (fig. 13-5) will be larger for one-half

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ycle of the input (one-half recurrence period) han for the other.

Assume that the terminal 2 arm of R249 ig. 13-18) is moved to the extreme right (as hown). Since the V61B grid is positive during he A period and negative during the B period, he V61B conduction through R249 and R256 uring the A period is greater than the conduc- on during the B period. Thus, the voltage at he terminal 2 arm of R249 will be more egative during the A period than during the B eriod. This increased negative potential is fed o the grids of r-f and i-f stages (V1, V3, and 4 of figure 13-5), and the gain of these stages uring the A period is less than the gain during he B period. Likewise, the amplitude of the ulse in the B period is increased, and the A eriod pulse amplitude is decreased.

By positioning the terminal 2 arm of R249 ig. 13-18) to the extreme left position (op- osite to that shown), the reverse effect to that ust discussed can be obtained. Since the input f V61A is negative during the A period and ositive during the B period, the conduction of 61A is smaller for the A period than for the B period. Thus, the negative voltages at the erminal 2 arm of R249 will be smaller for the A period than for the B period, and the gain of he r-f and i-f stages is greater during the A eriod than during the B period.

When the proper setting of R249 is reached, he gain of the video during the A period (master ulse) is equal to the gain of the video during the B period (slave pulse).

The gain control, R256 (fig. 13-18) is used to djust the magnitude of the negative voltage fed to he r-f and i-f stages for the various settings of 249. If the control is turned clockwise (upward) he negative potential fed to the r-f and i-f rids for any settings of R249 is decreased. onversely, if the gain control, R256, is turned counterclockwise (downward), the negative potential to the r-f and i-f grids for any setting of R249 will be increased.

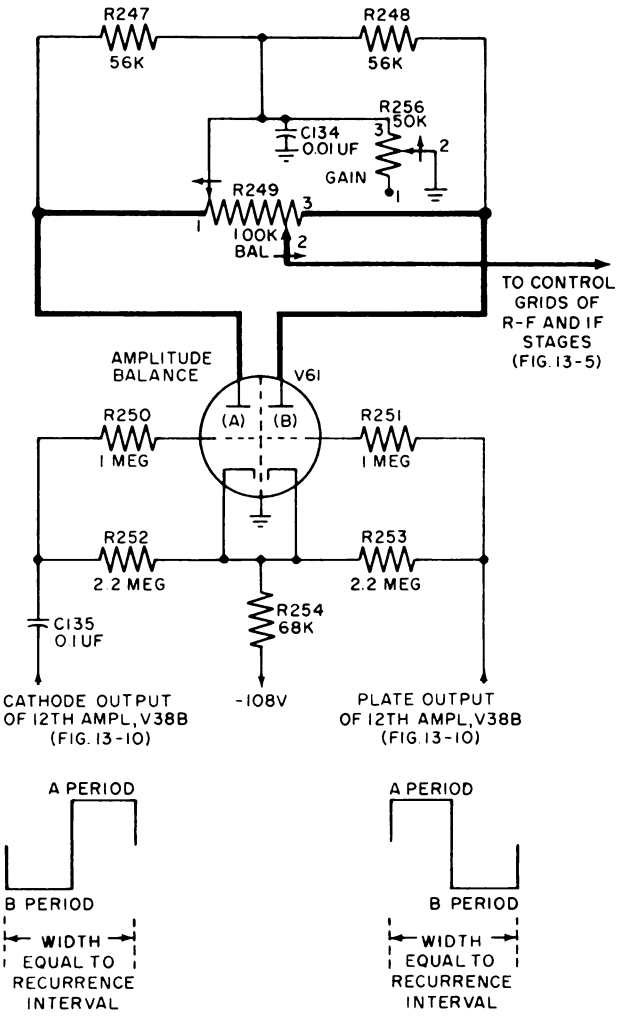

Figure 13-18.—Amplitude balance circuit.

QUIZ

1. The receiving set of the loran system (fig. 13-1) is basically an instrument for making what type measurements?
2. Identification of loran transmitting stations is accomplished in what manner?
3. The channels used in the loran system operate on what frequencies?
4. To properly select a given pair of transmitting stations, what controls (fig. 13-1) must be operated?
5. The input signal level to the r-f amplifier stage (fig. 13-2) is controlled by what switch?
6. An antijam switch, S4 (fig. 13-2) functions to reduce the effects of jamming by removing what components of the video signals?
7. The frequency control circuits (fig. 13-2) synchronize the repetition rate of the receiving set with what other source?

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8. Either automatic or manual frequency control may be used in the loran receiving system depending upon the setting of what switch (fig. 13-2)?
9. When AFC is used, the repetition rate of the incoming signals is compared with the receiving set repetition rate in what stages (fig. 13-2)?
10. The time base of the receiving set is produced by what stages (fig. 13-2)?
11. Frequency division is used in the counter stages, V20 through V31 (fig. 13-2) to produce a square wave at each counter output that bears what relationship to the counter input?
12. The square wave at the plates of the third to eleventh counters (V22 through V30 of figure 13-2) are employed in what circuits?
13. In the pulse repetition rate circuits (fig. 13-2) the input square waves from the third to eleventh counters (V22 through V30) are combined in the reset thyratron, V37, to produce a positive and negative reset pulse at what time?
14. The recurrence frequency of the reset pulse from V37 (fig. 13-2) bears what relationship to the recurrence frequency of the received signals?
15. The fixed and variable delay stages (fig. 13-2) combine their various inputs to produce a trigger pulse, which is coupled to what circuits?
16. Which of the pedestal pulses is variable from a time shortly after the reset pulse initiating the B period (see V37 in figure 13-2) to the end of the available delay time?
17. In sweep function 1 the pedestal pulses from V55 are displayed on the indicator tube, V60, in what manner?
18. The multiple taps on the inductance, L1 (fig. 13-4) provide what feature of the antenna coupler?
19. In the LOCAL and INTERMEDIATE positions of S2 (fig. 13-5) T-type resistor network attenuators are used to introduce attentuations of approximately what values?
20. The cathode and first two grids of V2 (fig. 13-5) are used in what type circuit?
21. Video amplifier V7 (fig. 13-5) is essentially what type circuit?
22. Frequency control is accomplished by variation of the grid bias on what stage (fig. 13-6)?
23. The discriminator, V15 (fig. 13-7) utilizes the gated outputs of the A and B circuits (comprising V8 through V14) to produce a d-c output voltage, the output polarity of which depends upon what condition?
24. The discriminator (V15) output is applied to the control grid of what stage (fig. 13-6)?
25. The ringer and clipper stages, V18 and V19 (fig. 13-9) are included in the pulse-forming circuits to produce the sharp, negative driving pulses that are required to drive what circuits?
26. The point in time at which the reset pulse occurs (fig. 13-10) corresponds with the trailing edge of which of the negative pulses from the counter chain?
27. The delay crank (fig. 13-1) controls the position of what pedestal?
28. The clipper stage, V48, functions similarly to what type circuit?
29. The trigger pulse used to initiate the master pedestal (fig. 13-15) is produced in what circuits?
30. The blanking d-c restorer, V59 (fig. 13-16) receives its input trigger from what stage?
31. The amplitude balance stage, V61 (fig. 13-18) provides a means of equalizing what two signals?

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CHAPTER 14

ELECTRONIC AIDS TO NAVIGATION: TACAN (PART I)

INTRODUCTION

This chapter presents a block diagram discussion of Radio Set AN/URN-3. Particular emphasis is placed on the purpose and relationship of its component parts.

GENERAL DESCRIPTION

Radio Set AN/URN-3 is used at shore activities or aboard ship to provide the radio beacon or ground portion of an air navigation system called TACAN (Tactical Air Navigation). The AN/URN-3, its associated antenna groups and accessories, and Radio Set AN/ARN-21 (located in the aircraft) together make up a complete air navigation system through which an aircraft equipped with the AN/ARN-21 equipment can accurately determine its position. Figures 14-1 and 14-2 show the major components of shipboard and shore installations, respectively. The antenna groups for shore and shipboard installations differ in several ways, which will be discussed later in this chapter.

The AN/URN-3 comprises three major groups: the receiver-transmitter, antenna (shore or shipboard), and power supply test set. The specific functions of these equipments will be discussed in detail later.

As many as 100 aircraft may simultaneously obtain navigational information in conjunction with a single installation of the AN/URN-3. The set is capable of receiving on any one of 126 frequencies in the range of 1025 to 1150 mc. Transmission of information also takes place on 126 frequencies in the ranges of 962 to 1024 mc and 1151 to 1213 mc.

Two types of antennas are available for use. Each antenna operates on 63 channels, corresponding to low-band frequencies and high-band frequencies, respectively. Low band installations transmit at frequencies between 962 and 1024 mc inclusive, and receive at frequencies between 1025 and 1087 mc. High band installations receive in the range of 1088 to 1150 mc, and transmit in the range of 1151 to 1213 mc.

Two frequencies are used in each channel: one for receiving and one for transmitting. The frequency used for receiving in low band installations is 63 mc above the frequency used for transmitting in the same channel. In high band installations, the receiving frequency is 63 mc below the transmitting frequency.

The radio beacon (AN/URN-3) output consists of the beacon identification call, distance information signals, bearing information signals, and random pulses used to make up a constant duty cycle. The radio beacon periodically transmits its identifying call in International Morse Code, thus enabling the interrogating aircraft to determine the radio beacon with which it is in contact. The characters of the code consist of a train of pulse-pairs generated at a fixed rate of 1350 cps in the coder-indicator (discussed later). A mechanical keyer in the same unit accomplishes the coding. The aircraft receiver detects these regularly occurring pulse-pairs and reproduces the code as a keyed 1350 cps audio tone.

The identification call has priority over the distance information signals. The bearing information reference bursts (discussed later) have priority over the identification call. The relative duration of these signals are such that there is effectively no interruption of distance information or identification call.

An introduction to the basic principles of TACAN is given in Chapter 1 of *Electronics Technician 3*, NavPers 10188-A. It is assumed that you have read this material.

The aircraft radio set transmits distance interrogation signals and receives replies from the radio beacon. Under normal circumstances the range over which aircraft can obtain reliable distance information from the beacon is about

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:00 mi. The received information is presented .s polar coordinate data (fig. 14-3), having both iistance (in nautical miles) and azimuth (in egrees of magnetic bearing) of the aircraft rom the radio beacon.

The aircraft radio set transmits pairs of ulses on the assigned channel. The pulses of ach pair are spaced 12 μs apart. The radio beacon receives the distance interrogation pulse-pairs, and, in reply, transmits back to the aircraft pulse-pairs on the assigned channel.

The time delay between the distance interrogation pulse-pair (transmitted from the aircraft) and the corresponding distance pulse-pair reply (transmitted from the radio beacon) is adjusted at the radio beacon to exactly 50 μs. The aircraft radio set deducts 50 μs from the

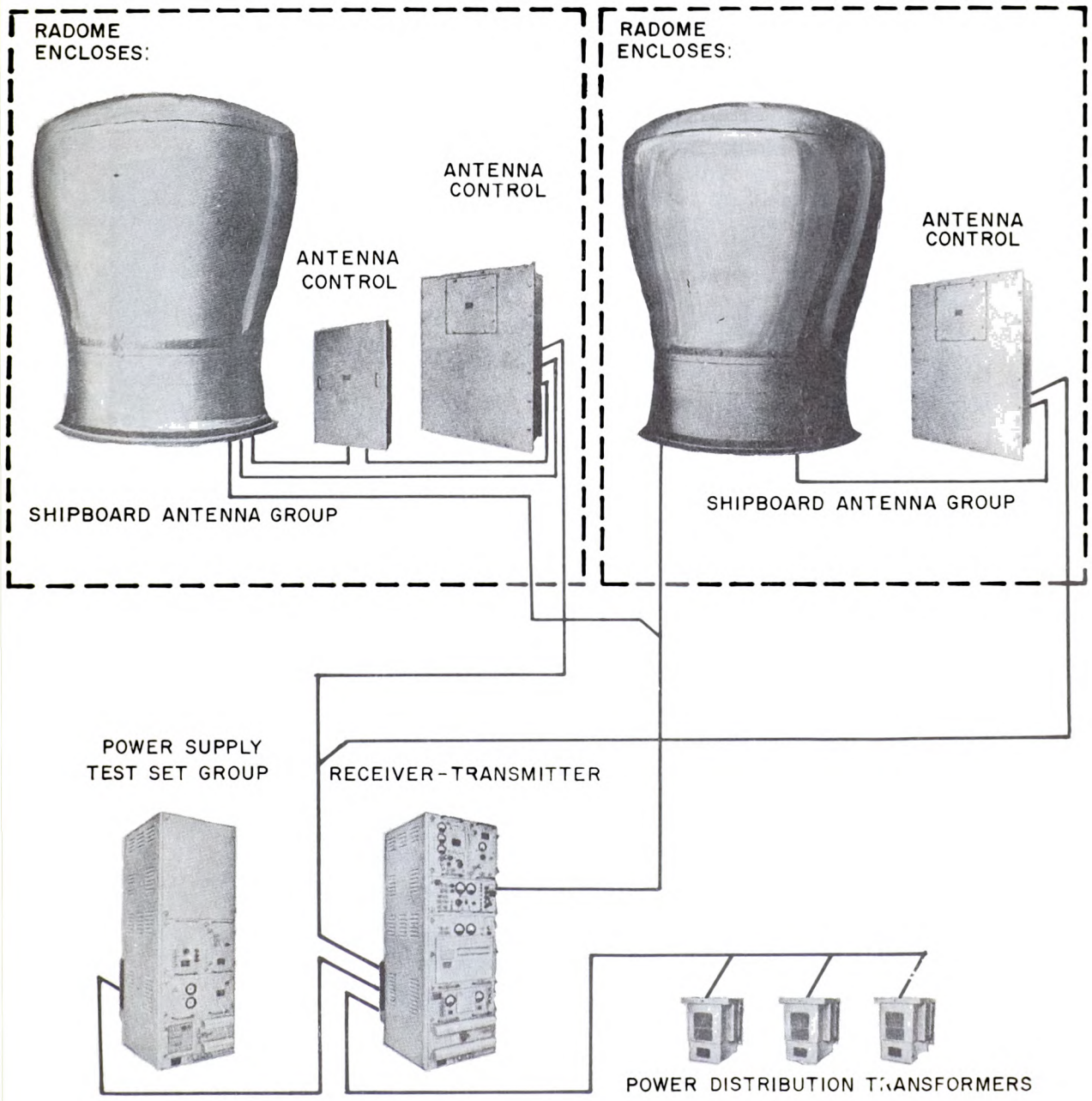

Figure 14-1.—Major components of Radio Beacon Set, AN/URN-3.

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total time elapsed between the interrogation and reply. The distance between the aircraft and the radio beacon is thus determined by measuring the total time elapsed between initial transmission of the distance interrogation pulse-pair and reception of the corresponding radio beacon reply.

The sequence of pulse-pairs transmitted by the AN/ARN-21 (aircraft equipment) is at a frequency of the radio beacon, and at a random rate peculiar to the particular equipment. When receiving, the AN/ARN-21 selects its own particular distance-measuring reply from all other signals by comparing the repetition rate of the reply pulses with the repetition rate of those originally transmitted. The receiver locks in on the reply pulses whose repetition rate is found to be the same as the rate of its interrogation pulses.

Bearing information originates from the antenna within the radio beacon. A single stationary vertical antenna (fig. 14-4, A) provides an omnidirectional pattern, such as shown in figure 14-4, B. A 15-cps parasitic element is properly spaced from the antenna so as to deflect the r-f energy to produce the cardioid

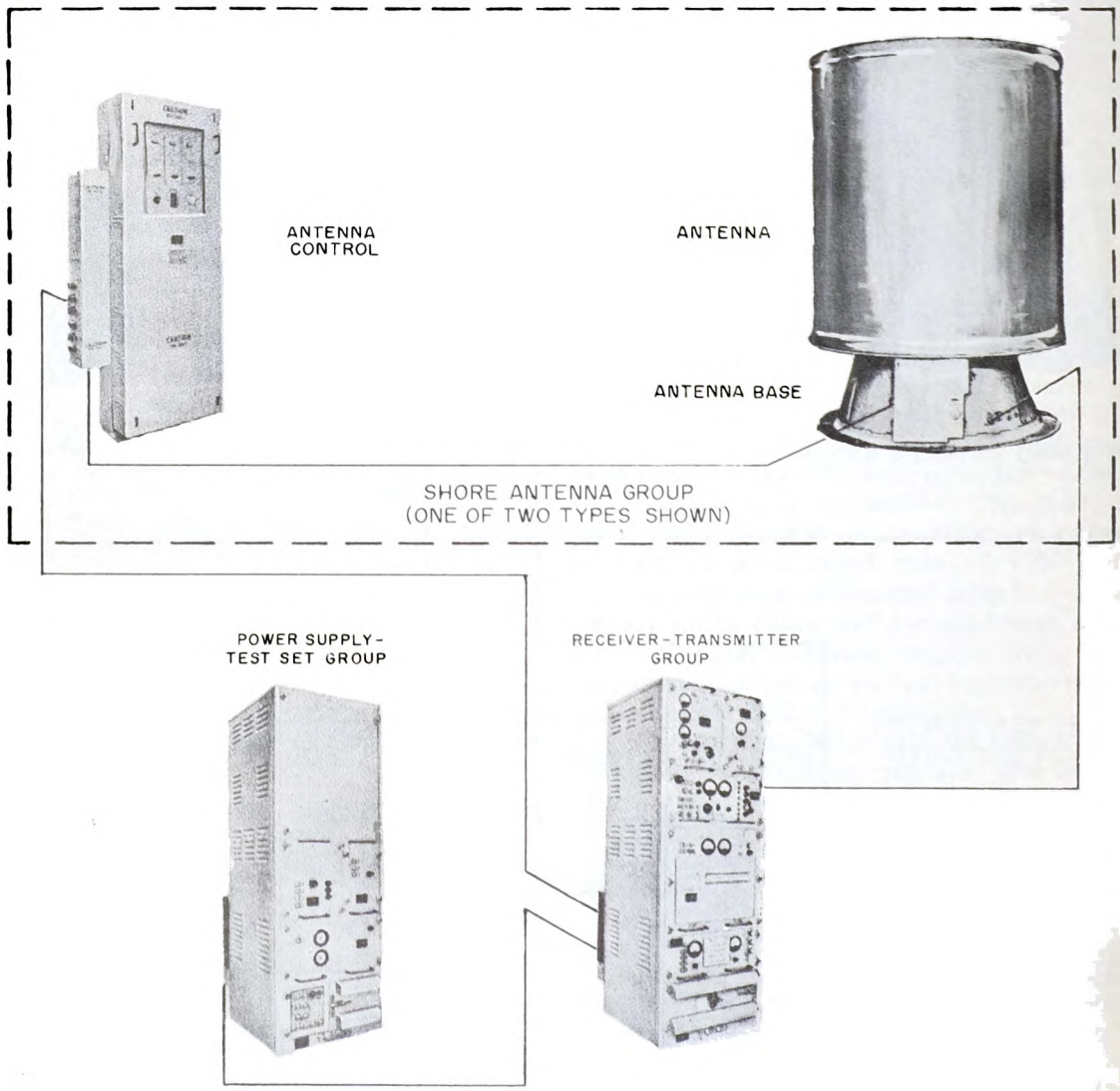

Figure 14-2.—Major components of Radio Beacon AN/URN-3 with associated shore antennas.

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attern of figure 14-4, C. The parasitic element s motor driven so that it encircles the antenna at rate of 15 rps. Thus, the parasitic element of ne antenna modulates the total radio beacon out-ut with a subaudio (15 cps) frequency component.

Encircling the 15-cps antenna element is an rrangement of eight vertical reflectors em-edded in an outer fiberglass cylinder, which is lso rotated at the 15-rps rate. Mechanically nked to each of the rotating cylinders, and nounted horizontally below the central array is n assembly referred to as the pulser plate ig. 14-5, A). On the upper edge of the pulser late (fig. 14-5, B) there is one soft iron slug. n the lower edge of the circular plate there re eight similar slugs. As the pulser plate otates, the upper and lower slugs of the plate ass through air gaps of pickup coils. As the lug passes through the air gap of the pickup oil, the net reluctance of the magnetic circuit hanges, and a pulse is induced in the coil wind-ng. One coil is excited by the upper slug (15 ps) once for each revolution. A second coil is xcited by each one of the eight lower (auxiliary) lugs, in turn, at 40-degree intervals, except for ne 80-degree interval when the space is oc-upied by the pulse from the upper slug. The ulse from the upper slug occurs at a time be-ween the eighth and first pulse of the outer ylinder so that a modulation of 135 cps is roduced.

The pulse produced by the upper slug on the ulser plate is used to establish magnetic north s a reference point in time. The pulse triggers

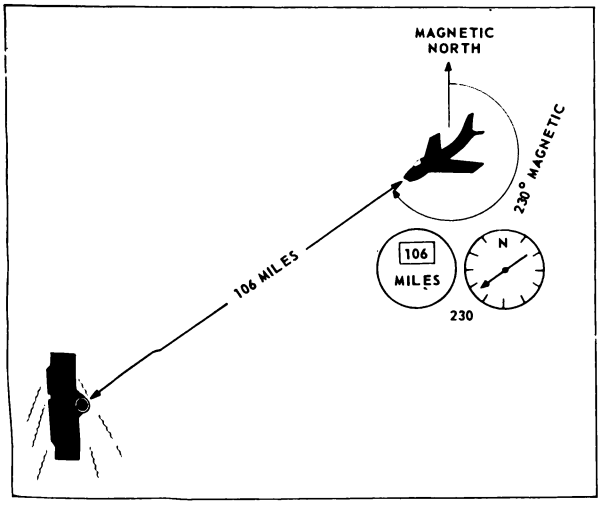

Figure 14-3.—Polar coordinate presentation of tacan data.

the generation of a coded burst, which is introduced into the 15-cps, a-m field of energy radiated by the beacon antenna. As stated above, the 15-cps pulse is generated once during each revolution of the reflector. The pulser plate causes the generation of the 15-cps trigger pulse each time the peak of the radiation lobe points due east (an arbitrary point). The pulse thus obtained causes the generation of the north reference burst.

The aircraft receiver utilizes the 15-cps and 135-cps modulation on the total received signal (and the 15-cps and 135-cps reference burst, which are also components of the received signal) in presenting azimuth information. Figure 14-6 shows the position of the north reference burst on the 15-cps sine wave (detected from the a-m, r-f carrier) as seen by four aircraft at four different points around the radio beacon. Figure 14-6, A, shows the position of the north reference burst on the 15-cps modulation for an aircraft due south (magnetic) of the beacon.

To understand this action, first consider the direction of rotation of the pattern and the signal components that the aircraft will receive. The pattern is rotating in a clockwise direction. With the pattern as shown at A, the aircraft will receive a portion of the total signal strength, which is neither maximum nor minimum. If the plane were repositioned to a point "a," minimum signal strength would be received. At "b" the signal strength received is again neither maximum nor minimum. In the latter case, however, the signal strength would be moving in the negative direction because the positions immediately following that shown at "a" would yield still a smaller signal amplitude at the aircraft. At "c," an aircraft would receive maximum signal strength from the radio beacon. Thus the signal received by the aircraft appears as a 15-cps sine wave with the position of the beacon represented by a vertical line (north burst) at some point along the X axis. The position of the waveform with respect to the vertical line varies with the aircraft position. The north reference burst always occurs when the cardioid pattern points due east.

In figure 14-6, B, the aircraft is receiving minimum signal amplitude at the time of the north reference burst. Its bearing relative to the ground radio beacon is 270° (magnetic). In figure 14-6, C, the aircraft is approaching the beacon from a point due (magnetic) north. The signal received by the aircraft would be half

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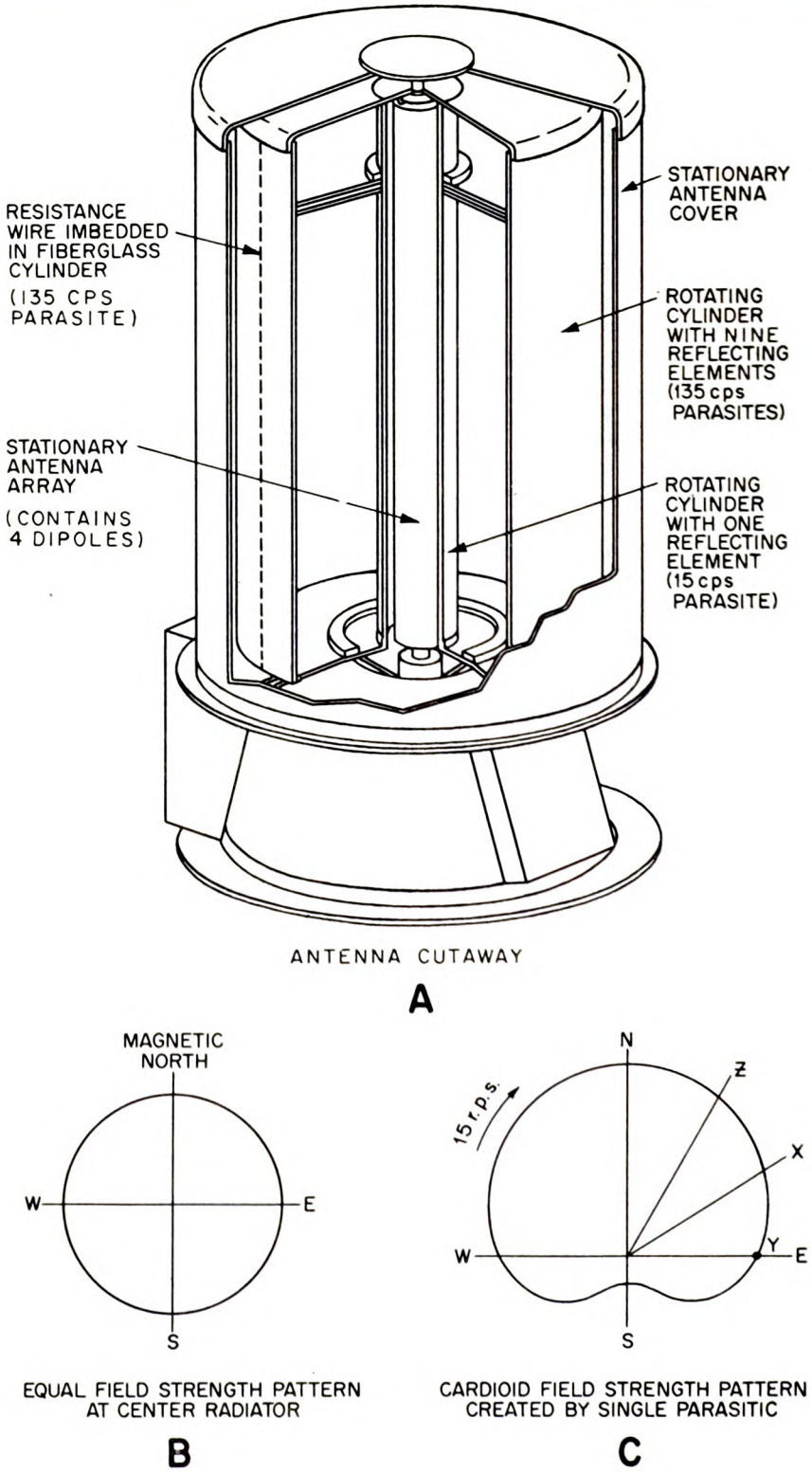

Figure 14-4.—Construction of the antenna and antenna patterns.

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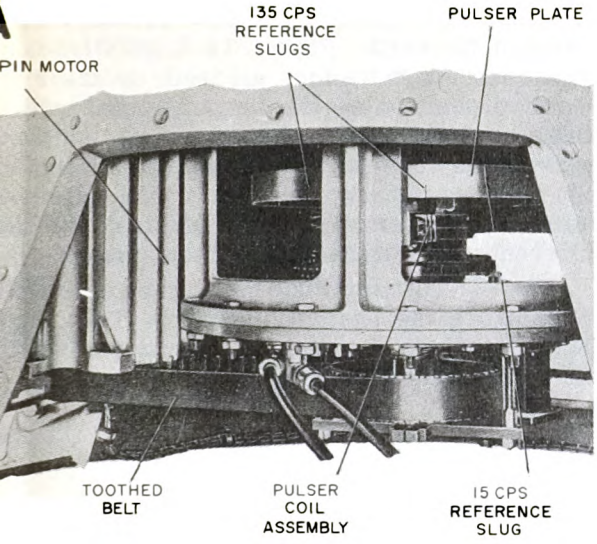

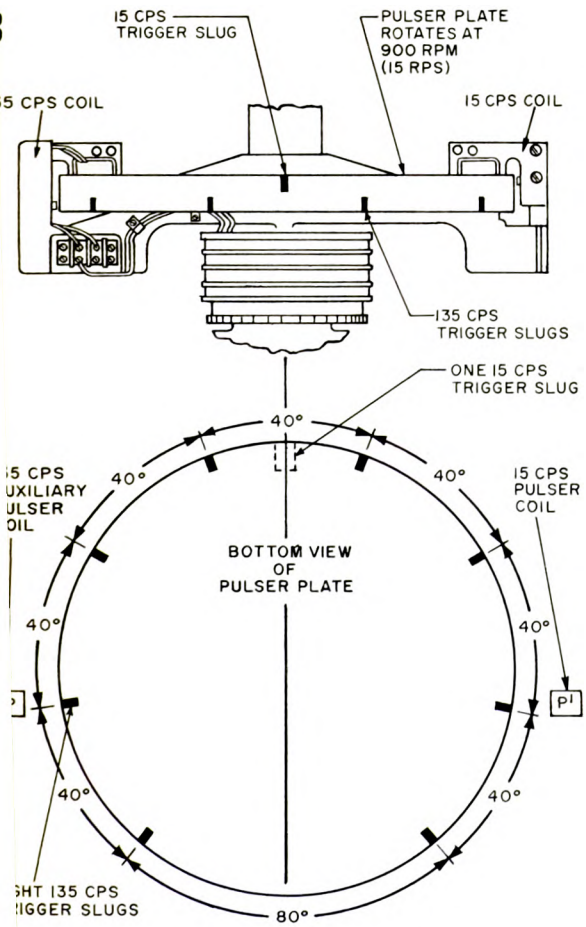

Figure 14-5.—Method of generating reference pulses.

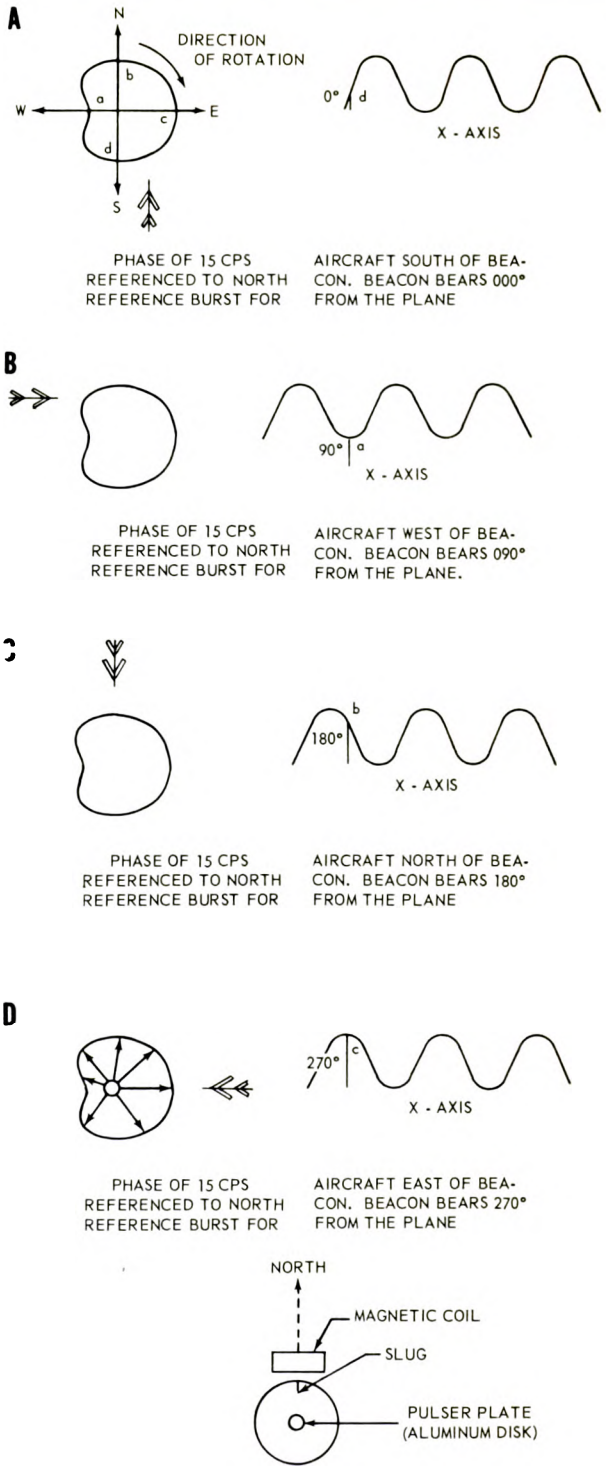

Figure 14-6.—Relative positions of the north reference burst.

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amplitude (on the 15-cps sine wave) and moving to the negative portion of the wave. The magnetic bearing of the beacon from the aircraft is therefore 180°. In figure 14-6, D, the aircraft is approaching from due east, and receiving maximum signal strength from the radio beacon. The north reference burst would therefore appear at the maximum point on the receiver sine wave, indicating that the beacon bears 270° magnetic.

The phase (time) difference between the 15-cps sine wave and the north reference burst is detected by the aircraft receiver of Radio Set AN/ARN-21 and converted to a meter reading in degrees, showing the magnetic bearing from the aircraft to the beacon with respect to magnetic north. Figure 14-7 illustrates the signal received by four aircraft (in their respective position as the beacon signal rotates through 360°) and the meter reading that would correspond to a particular bearing. Note that all the aircraft receive the north reference burst at the same instant of time but at a different phase of the 15-cps, a-m wave.

The 135-cps amplitude modulation is provided to improve the bearing accuracy. This signal is also imposed on the r-f energy radiated by the radio beacon antenna. As stated earlier, the

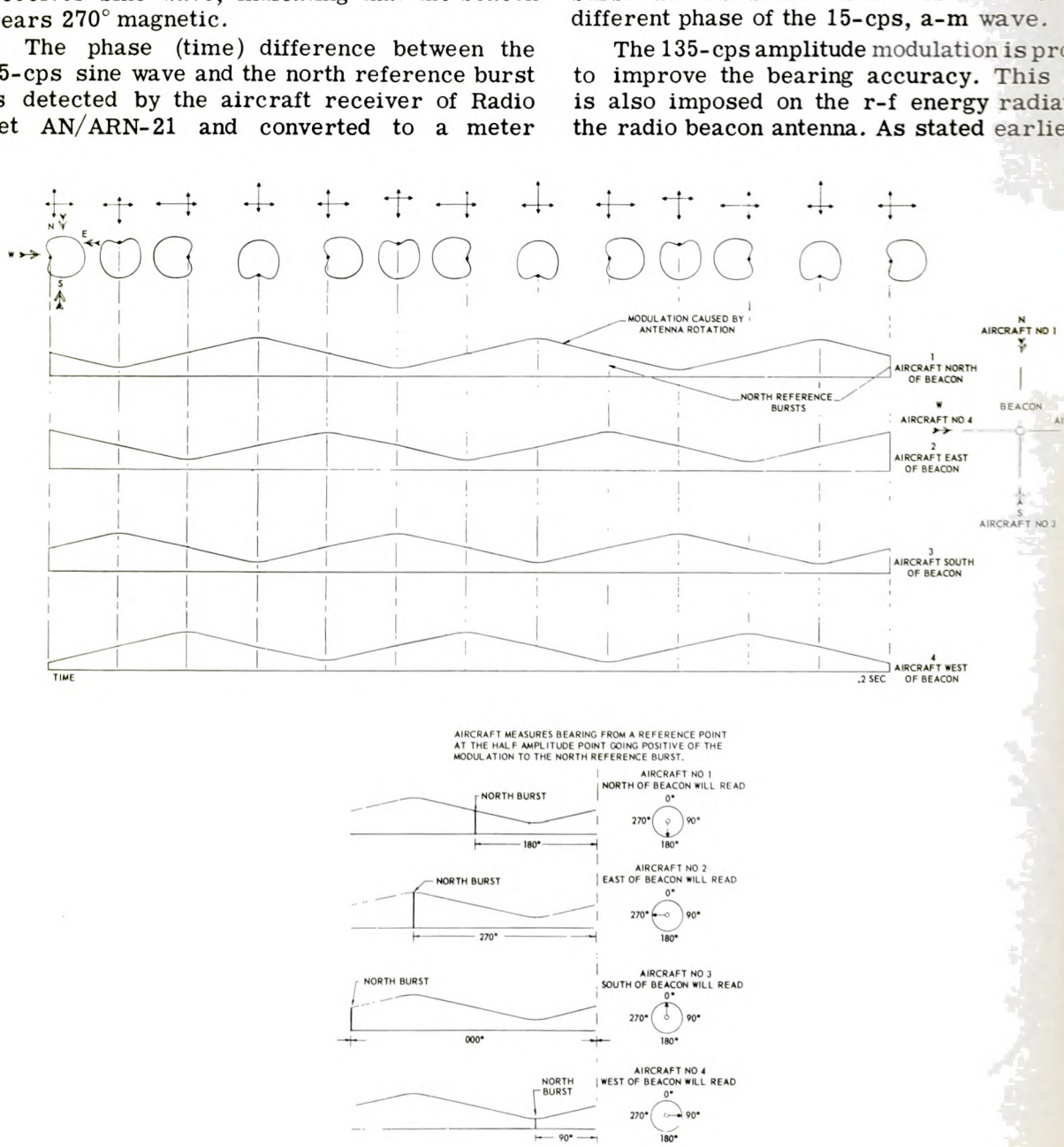

Figure 14-7.—Meter readings for various aircraft positions.

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35-cps signal is generated in the same method s the 15-cps component, with its directors paced 40° apart around the radiating elements, nd rotating about the antenna at a 15-rps rate.

The 135-cps signal alters the cardioid pattern o form a new pattern, as shown in figure 4-8. Thus, the antenna pattern basically consists of a large lobe (from the 15-cps director) nd nine smaller lobes (produced by both the 5-cps and the 135-cps directors).

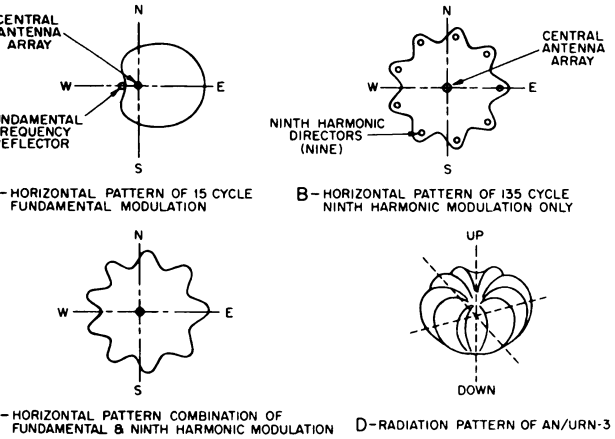

Figure 14-8.—Effects of 15 cps and 135 cps modulation.

Figure 14-9 summarizes the steps involved n the development of the composite 15-cps and 35-cps signal elements into the total bearing nformation supplied by the radio beacon to the ircraft. The signals, including the 15-cps and 35-cps amplitude modulation and the 15-cps and 35-cps reference bursts, received by the aircraft in the positions due east, south, west, and orth relative to the beacon, appear as shown. gain, the phase of the modulation relative to the urst depends on the position of the aircraft elative to the beacon. The 135-cps reference ursts, are used to refine or make more accurate the measurement of the angle relative to nagnetic north.

The total radio beacon output consists of the nternally generated reference-frequency, nodulated by pulse-pairs. The pulse-pairs are ransmitted at a rate of 3600 pulse-pairs per econd. Nine hundred of the total pulse-pairs er second are the 15-cps and 135-cps reference ursts. (The reference bursts are initiated by he 15-cps and 135-cps trigger pulses, respectively). The additional 2700 pulse-pairs occurring at a rate of 1350 cps (double pulsed rate) are random squitter pulses generated to maintain a constant duty cycle.

The composite radio beacon pattern is shown in figure 14-10. In 1/15 second (1 cycle of 15-cps signal), 12 north reference burst pulse-pairs occur. Over a one-second period, 12 X 15 pulse-pairs occur, which total 180 north reference bursts per second. Seven hundred and twenty auxiliary (135 cps) reference burst pulse-pairs occur each second. These pairs are divided into 6 pulse-pairs occurring 120 times per second (8 times per cycle). Random pulse-pairs (totaling 2700 pulse-pairs) are inserted between the reference bursts. Identity pulse-pairs are substituted for the random pulse-pairs once every 37.5 seconds.

Of all the signals transmitted by the radio beacon, distance-measuring information alone is not generated within the beacon. Distance reply signals are supplied by the radio beacon only when interrogated by the airborne radio set. The AN/ARN-21 sends out a sequence of paired pulses at a frequency of the radio beacon receiver and at a random rate peculiar to the particular equipment. This signal is received by the radio beacon receiver and re-transmitted to the aircraft, as described earlier. All signals transmitted by the radio beacon are characterized by the fact that they consist of pulse-pairs, with 12-μs spacing between the two pulses of the pair. The number of pulse-pairs per second and the spacing between the pulse-pairs (the spacing between the leading edge of the first pulse-pair and the leading edge of the first pulse of the next pair) depend upon the particular signal element and is a characteristic of that particular signal only. However, it is the spacing of 12 μs between the pulses of a pair, which provides the aircraft radio set with the means of distinguishing between the signal pulses from the radio beacon and other pulses that may be present on the received radio frequency.

The three elements of information (bearing, identification call, and distance measuring information) are transmitted by the radio beacon as a train of pulse-pairs. To prevent interference between the three signal elements, the radio beacon is equipped to assign a sequence of priority by which each element enters into the over-all pulse train.

First in order of priority are the bearing reference bursts. Because they occur at a fixed rate, they are of relatively short duration and

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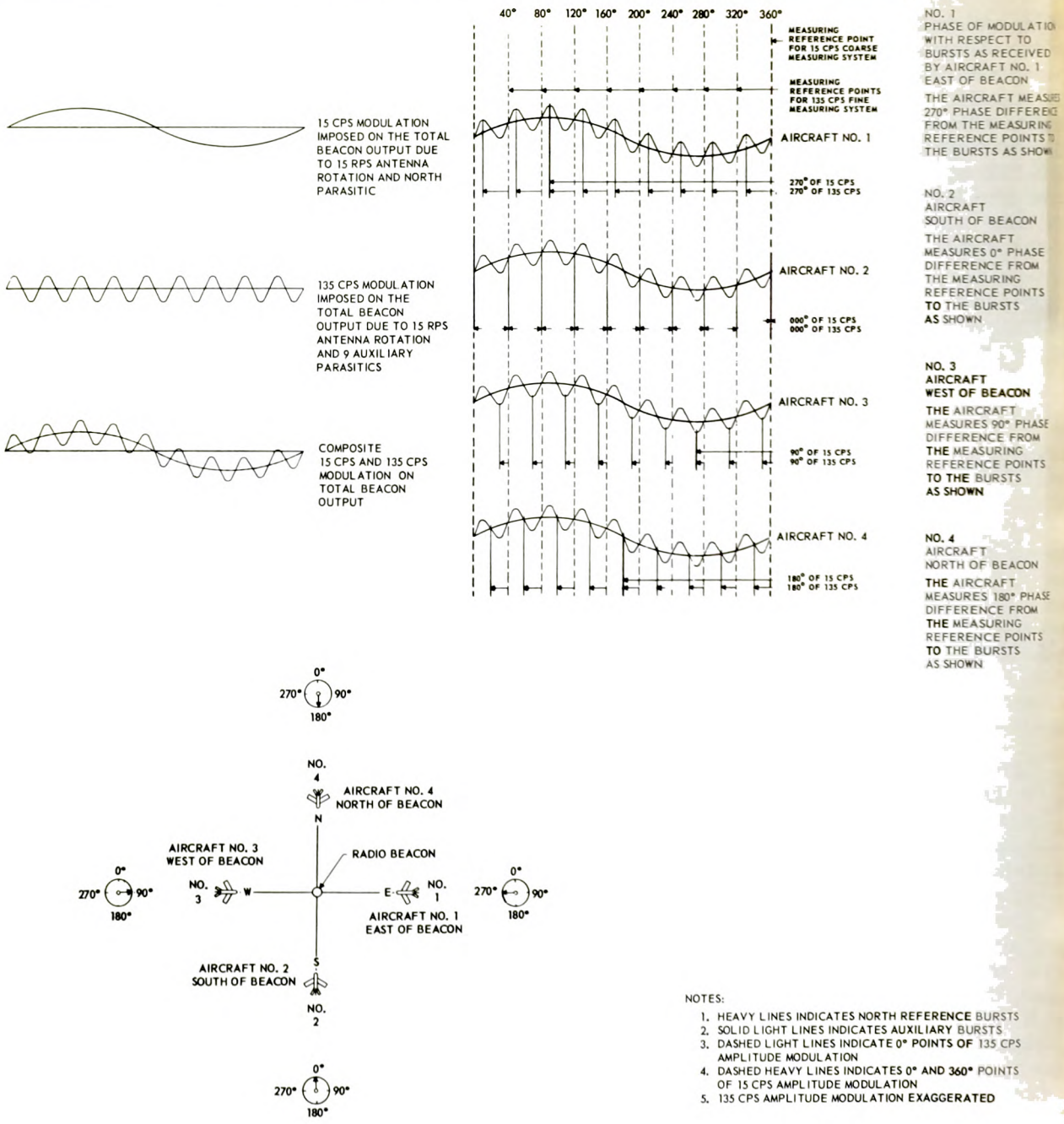
15 CPS MODULATION IMPOSED ON THE TOTAL BEACON OUTPUT DUE TO 15 RPS ANTENNA ROTATION AND NORTH PARASITIC
135 CPS MODULATION IMPOSED ON THE TOTAL BEACON OUTPUT DUE TO 15 RPS ANTENNA ROTATION AND 9 AUXILIARY PARASITICS
COMPOSITE 15 CPS AND 135 CPS MODULATION ON TOTAL BEACON OUTPUT
40° 80° 120° 160° 200° 240° 280° 320° 360°
MEASURING REFERENCE POINT FOR 15 CPS COARSE MEASURING SYSTEM
MEASURING REFERENCE POINTS FOR 135 CPS FINE MEASURING SYSTEM
AIRCRAFT NO. 1
270° OF 15 CPS
270° OF 135 CPS
AIRCRAFT NO. 2
000° OF 15 CPS
000° OF 135 CPS
AIRCRAFT NO. 3
90° OF 15 CPS
90° OF 135 CPS
AIRCRAFT NO. 4
180° OF 15 CPS
180° OF 135 CPS
NO. 1
PHASE OF MODULATION WITH RESPECT TO BURSTS AS RECEIVED BY AIRCRAFT NO. 1 EAST OF BEACON
THE AIRCRAFT MEASURES 270° PHASE DIFFERENCE FROM THE MEASURING REFERENCE POINTS TO THE BURSTS AS SHOWN
NO. 2
AIRCRAFT SOUTH OF BEACON
THE AIRCRAFT MEASURES 0° PHASE DIFFERENCE FROM THE MEASURING REFERENCE POINTS TO THE BURSTS AS SHOWN
NO. 3
AIRCRAFT WEST OF BEACON
THE AIRCRAFT MEASURES 90° PHASE DIFFERENCE FROM THE MEASURING REFERENCE POINTS TO THE BURSTS AS SHOWN
NO. 4
AIRCRAFT NORTH OF BEACON
THE AIRCRAFT MEASURES 180° PHASE DIFFERENCE FROM THE MEASURING REFERENCE POINTS TO THE BURSTS AS SHOWN
0°
270° 90°
180°
NO. 4
AIRCRAFT NO. 4 NORTH OF BEACON
N
AIRCRAFT NO. 3 WEST OF BEACON
RADIO BEACON
NO. 3
W
E
NO. 1
AIRCRAFT NO. 1 EAST OF BEACON
S
AIRCRAFT NO. 2 SOUTH OF BEACON
NO. 2
NOTES:
1. HEAVY LINES INDICATES NORTH REFERENCE BURSTS
2. SOLID LIGHT LINES INDICATES AUXILIARY BURSTS
3. DASHED LIGHT LINES INDICATE 0° POINTS OF 135 CPS AMPLITUDE MODULATION
4. DASHED HEAVY LINES INDICATES 0° AND 360° POINTS OF 15 CPS AMPLITUDE MODULATION
5. 135 CPS AMPLITUDE MODULATION EXAGGERATED

Figure 14-9.—Converting 15 cps and 135 cps amplitude modulation and burst signal elements into bearing information.

convey the most important portion of the information supplied by the beacon. Accordingly, reference bursts will be transmitted regardless of the number of interrogations.

Second in order of priority is the identity call signal. The identity call signal also occurs at a fixed rate. However, the duration of the identity call is sufficiently long, so that the utilization of

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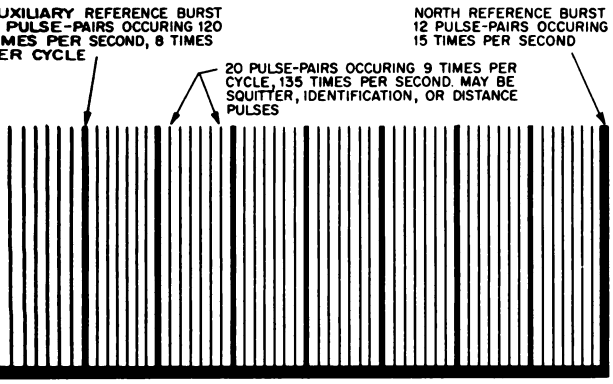
UXILIARY REFERENCE BURST
PULSE-PAIRS OCCURING 120
MES PER SECOND, 8 TIMES
ER CYCLE
NORTH REFERENCE BURST
12 PULSE-PAIRS OCCURING
15 TIMES PER SECOND
20 PULSE-PAIRS OCCURING 9 TIMES PER CYCLE, 135 TIMES PER SECOND. MAY BE SQUITTER, IDENTIFICATION, OR DISTANCE PULSES

Figure 14-10.—Composite radio beacon signal.

ne portion of the total duty cycle required for ne insertion of the bearing reference burst does ot materially affect the reception of intelligible dentity code by the aircraft.

Third in order of priority are the distance nterrogation signals. The interrogation pulses re not initiated within the beacon, and their rrival within the pulse train must, therefore, e controlled to prevent their interfering with ne other signal elements of the radio beacon. his is done by allowing the interrogation pulses o enter the pulse train only during a time nterval not occupied by the reference burst or he identity call signals. The time occupied by ne reference burst is small in comparison with ne time between them. A memory circuit built nto the aircraft receiver makes it possible to ompensate for the time interval occupied by the dentity pulses so that no apparent interruption f distance-measuring signals results.

Last in order of priority are "squitter" ulses consisting of random noise pulses, which re used as fill-in signals when the maximum umber of aircraft are not interrogating the eacon. This is necessary to maintain the nodulation pattern and to permit a constant luty cycle.

BLOCK DIAGRAM

RECEIVER

The functional arrangement of the major cir-uits of the radio beacon is illustrated in figure 4-11. The antenna receives distance informa-ion pulse-pairs from the aircraft and transmits ack to the aircraft the total beacon signals, ncluding the distance-measuring reply pulse-airs. The antenna also generates north and auxiliary trigger pulses at rates of 15 and 135 pps, respectively. These trigger pulses are generated in separate pulser coils, as described earlier (see figure 14-5). The 15- and 135-pps antenna trigger pulses initiate the generation of bearing information.

A single antenna is used for both transmission and reception. This is made possible through the use of the antenna duplexer. Distance interrogation pulse-pairs are received by the antenna and are passed through the duplexer to the radio receiver. The output of the transmitter section is routed through the duplexer and transmitted from the antenna. The methods employed to prevent the entrance of any portion of the beacon transmitted signal to the beacon receiver will be discussed later.

A single crystal oscillator, V1, provides the basic frequency for the transmitter and the heterodyne frequency for the receiver. The receiver multiplies this frequency 24 times in 5 stages (V1 through V5) of frequency multiplication. The final heterodyne frequency derived through this process is applied to a "balanced Tee" mixer stage comprising CR1 and CR2. The receiver local oscillator and receiver r-f input signals are always separated by 63 mc.

The receiver r-f input signal contains the distance information and random noise signals, which are heterodyned in the "balanced Tee" mixer stage with the local oscillator signal to produce a 63-mc intermediate frequency. The output of the mixer stage (containing distance interrogation pulse-pairs with 12-μs spacing between the pulses of a pair, and random noise pulses with random spacing) is applied, in turn, to three stages of preamplification (V6 through V8). The signal output of the preamplifiers is fed through five stages of i-f amplification (V9 through V13). The preamplifier and i-f amplifier stages raise the signal to a level suitable for detection.

A Ferris discriminator, V14, employing both wide and narrow band detection, eliminates adjacent channel interference, and provides a narrow band of frequencies at its output. Wide band signals are degenerated. The output of the Ferris discriminator is a video signal.

The detected pulses from the discriminator are amplified in V15 through V17 and fed through a gating circuit V18 (discussed later) to another pulse amplifier, V19, and thence through a delay line, W1, to a coincidence decoder circuit V20. The coincidence decoder, V20, passes only the second of a pair of pulses spaced 12 μs apart.

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Proper spacing is determined by comparing the time of arrival of the first pulse of a pair passed through a 12-μs delay line, V1, with the time of arrival of the undelayed second pulse of the pair. The first incoming pulse is delayed so that it acts as a gate to pass the second undelayed pulse. Only if the second pulse follows the first by 12μs will it be able to pass through the gate. Thus, an acceptably paired pulse (one with 12-μs spacing between the pulses of a pair) yields a single pulse at the output of the decoder, V20. No output pulse results from any other combination of pulses.

A one-shot multivibrator, V21, triggered by the output of the 12-μs coincidence decoder, V20, provides receiver output pulses of uniform shape and amplitude. The V21 multivibrator has three outputs. One of the outputs of the multivibrator triggers a 40-μs blanking circuit comprising V27 and V28. As a result of this pulse, the receiver is blanked out for 40 μs after each decoded pulse-pair to insure sufficient recovery time in the coder-indicator and the units that constitute the transmitter, and, to insure a spacing of at least 40 μs between the decoded random pulses. A pulse from the transmitter low-level modulator (discussed later) is also fed to the blanking circuit to blank the receiver during transmission. This prevents the beacon from being interrogated by its own transmitter pulses.

The second output of the multivibrator, V21, feeds a squitter circuit comprising V24 through V26. The squitter circuit functions to produce a negative control voltage, the magnitude of which depends on the number of pulses passing through the multivibrator (during an interrogation period) to the receiver output. The negative control voltage is fed back to the i-f amplifiers (as bias) to control the gain of these stages. By controlling the gain of the i-f stages, the number of noise pulses of sufficient amplitude (and correct spacing) to trigger the multivibrator, V21, can also be controlled.

To maintain a constant radio beacon duty cycle of 3600 pulse-pairs per second, the output of the radio receiver must be held at 2700 pps, regardless of the number of distance interrogation pulses received. If the number of interrogations falls below 2700 pps, a sufficient number of random noise pulses is permitted to pass through the i-f circuits and are decoded to maintain the receiver output at 2700 pps. If the number of interrogations exceeds 2700 pps, a sufficient number of noise pulses is rejected to maintain a 2700-pps-output.

The third output of the multivibrator, V21, is fed to a cathode follower stage, V22. The output of this stage is the output of the receiver. This output, composed of distance interrogation pulses and decoded noise pulses totaling 2700 pps with a minimum spacing of 40μs between pulses, is fed to the keyer tube in the coder-indicator.

CODER-INDICATOR

The output of the radio receiver is applied to the keyer stage, V14, in the coder-indicator. A 1350-cps, double-pulsed audio signal is also applied to the keyer stage. The identification call keyer is a mechanical coding wheel (fig. 14-12) with a prearranged International Morse Code signal which identifies the transmitting radio beacon. The code is set by extending the metal segments (secured with screws) around its edge. As the wheel rotates, the extended segments depress the roller of the keying switch, which is in the transmitter keying circuit. When the keyer tube is keyed, the 1350-cps signal is allowed to pass. If the keyer tube is not keyed, only the signals from the radio receiver are allowed to pass.

The output of the keyer stage, V14 (fig. 14-11) is applied to a reference burst gating amplifier, V15. Whether or not the signal from the keyer stage will be allowed to pass through the reference burst amplifier at any particular moment depends upon the priority system established for the signal elements of the radio beacon.

The 15- and 135-cps reference bursts (which are required for the aircraft to determine its

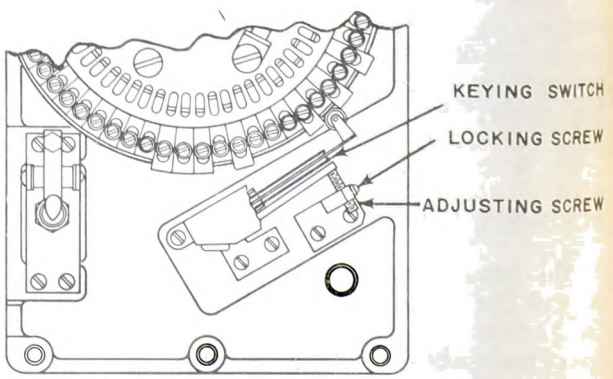

Figure 14-12.—Identification call keyer.

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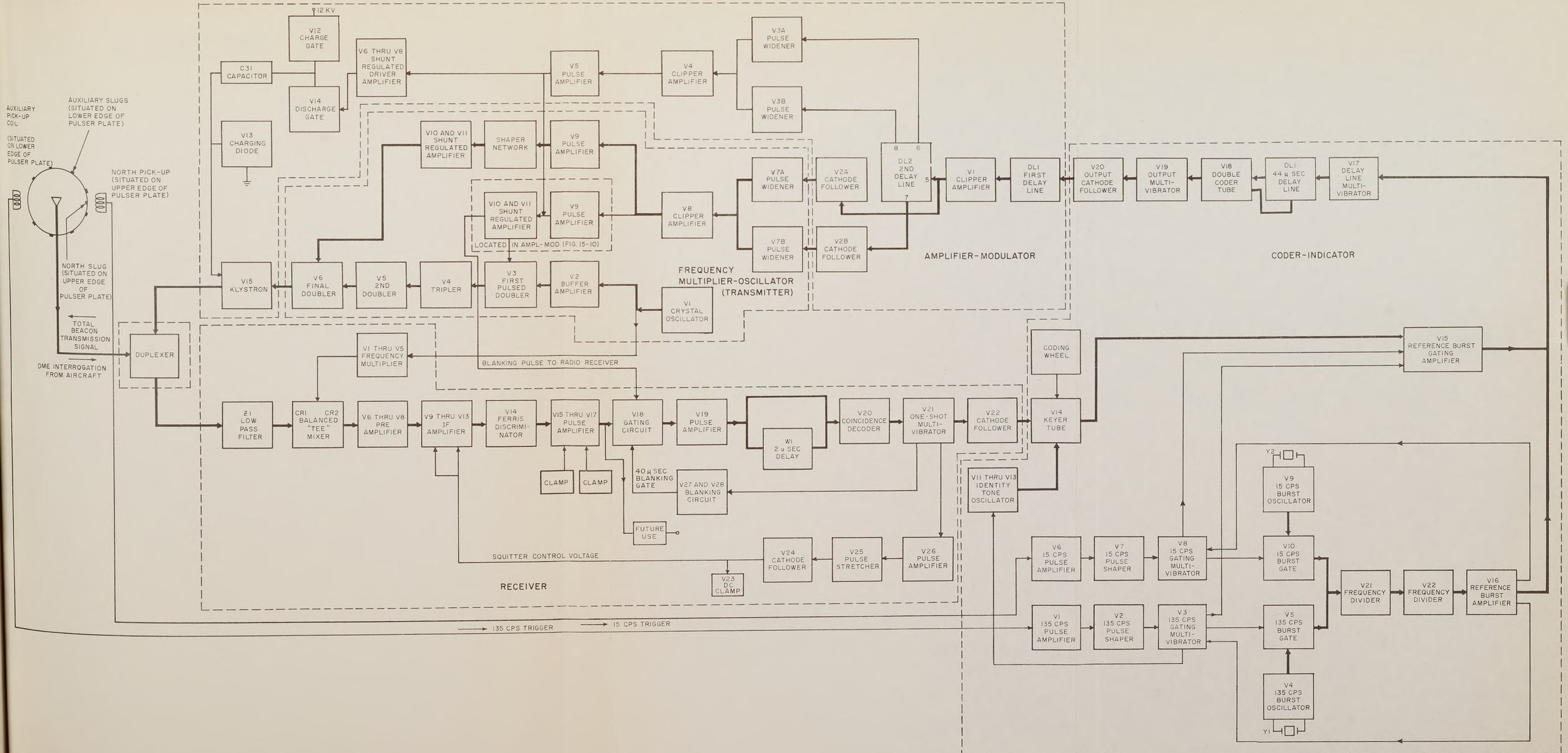

Figure 14-11.—Radio beacon block diagram

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aring) must be transmitted in preference to other radio beacon elements. Consequently, y signal from the radio receiver or identifica-n code system will be canceled whenever the ference burst signals are being transmitted. e reference burst gating amplifier functions to sign the orders of priority as required.

Bearing information is initiated by the antenna the radio beacon. The soft iron slugs and ckup coils (already discussed) are responsible initiating the 15- and 135-cps reference rsts. Each generated trigger pulse affects its sociated reference burst circuit to produce the rth and auxiliary reference burst pulses, spectively.

The north burst is triggered at a 15-cps te by the trigger generated in the upper tenna pulser coil. (See fig. 14-5, B.) The gger pulse is amplified and shaped by V6 d V7 to trigger the 15-cps gating multi-brator, V8. The multivibrator generates a ctangular positive pulse, which triggers the -cps burst gate stage, V10. The north burst lses originate in a crystal oscillator, V9. The cillator operates at four times the required petition rate to generate 133.33-kc sine waves 5 μs duration). The oscillator signals pass rough the burst gate stage, V10, under the fluence of the gating pulse from the multi-brator, V8. The duration of the gating pulse lows the passage of 48 oscillator sine waves. ne 48 sine waves are divided by two in two ages (V21 and V22) of frequency division.

The division circuits operate as flip-flop ultivibrators. The final output of the dividers nsists of 12 north reference burst pulses aced at 30-μs intervals, which are fed to the ference burst amplifier, V16.

Of special interest in the north burst gen-rating arrangement is the fact that the trigger nerated by the antenna pulser coil does not nerate the north burst directly. Instead, it is ed to generate a gate, which unblocks the 33.33-kc oscillator long enough for 48 oscil-tor output pulses to go through the frequency vider chain before a return trigger from the ference burst amplifier blocks the oscillator put again (by cutting off the multivibrator, g. 14-11), which prevents further generation of rth burst pulses.

The positive output of the 15-cps gating multi-brator, V8, is used to trigger the 15-cps rst gate as stated earlier. The negative rtion of the gating multivibrator output is plied (simultaneously) to the reference burst gating amplifier, V15, and is of sufficient negative amplitude to block or prevent the passage of signals from the keyer tube, V14, during periods of reference burst transmission.

The auxiliary bursts are triggered 120 times a second at a 135-cps rate. To understand this action, consider the trigger generating effect on the pulser coil, P (fig. 14-5, B) as the pulser plate turns through 360 degrees. Since the antenna is rotating at a 15-rps rate, the eight auxiliary slugs generate trigger pulses in P, totaling 120 pps (8 X 15) pulses per second). However, the slugs are spaced at only 40 degrees, which should normally allow 9 trigger pulses per second for the full 360° rotation. Thus, the repetition rate of the auxiliary pulses is that of 135 pps, while only 120 pulses are generated in the auxiliary pulser coil in one second. The ninth pulse (as stated earlier) is generated by the slug on the upper part of the pulser plate, and is placed at a center position between the first and eighth auxiliary slugs (40° from either slug).

The auxiliary bursts are produced in much the same way as the north bursts described earlier. The 135-cps burst oscillator (fig. 14-11) is crystal controlled by V4 to produce a frequency of 166.67 kc. The duration of one cycle of this output is 6 μs.

The gate from the 135-cps gating multivibrator, V3, permits the 135-cps burst gate, V5, to allow the passage of 24 cycles of the oscillator frequency. The resulting output is divided in two stages of frequency division (V21 and V22) to yield 6 output pulses, each of which has 24 μs-duration. The negative portion of the 135-cps multivibrator gate pulse is applied to the reference burst gating amplifier, V15, to block the identification coded and distance information signals as described earlier.

Note that the identity tone oscillator (V10 and V11) operates at the tenth multiple of the 135-cps gating multivibrator (1350). Consequently, a pulse from the 135-cps gating multivibrator, V3, is used to phase lock the identity tone oscillator.

The output of the reference burst amplifier, V16, alternating with the output of the reference burst gating amplifier, V15, is fed over a common path to the delay-line multivibrator, V17. This stage is connected as a one-shot multivibrator, which yields one uniform output pulse for each of the pulses received, whether the pulse originated in the reference burst amplifier or the reference burst gating amplifier.

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The 44-μs delay line, DL1, introduces part of the over-all required radio beacon delay of 50 μs to each of the pulses. In addition, the delay line causes two pulses to be applied to the double coder tube, V18, for each pulse applied to the line from the multivibrator, V17. Each of the pulse-pairs thus generated has a 12 μs spacing between them. The pulse-pairs are further shaped in the output multivibrator, V19, and fed to the output cathode follower stage, V20.

The output of the coder-indicator during one second consists of 3600 pulse-pairs, with a 12-μs spacing between pulses of a pair. Each of the pulses has a 15-volt, peak-to-peak amplitude, and a 1.5-μs pulse width.

AMPLIFIER-MODULATOR

The signals from the coder-indicator are applied to the first delay line, DL1, which adds sufficient delay to the 44-μs delay already discussed and stray delay through the beacon to bring the total delay to 50 μs. The first delay line output is fed through a clipper amplifier V1, to a second delay line, DL2.

The second delay line, DL2, feeds delayed pulses through cathode followers V2A and V2B to the pulse widener circuits, V7A and V7B, in the frequency multiplier-oscillator chassis. Another output of the second delay line, DL2, is applied to a similar pulse widener circuit (V3) in the amplifier-modulator chassis.

The pulse widener stages, V3A and V3B, of the amplifier-modulator receives its input pulses from taps on the delay line. This circuit requires two positive trigger pulses to produce a single output. The leading edge of the first delayed input (from the delay line) is caused to form the leading edge of the positive output pulse from the pulse widener circuit. Likewise, the leading edge of the second delayed input triggers the pulse widener circuit to form the trailing edge of the output pulse.

The width of the pulse from the pulse widener circuit is determined by the distance (in time) between the two delayed pulses from the second delay line. The pulse width (and pulse delay with respect to the time of the input) can therefore be varied by changing the position of the taps on the delay line. It might also be noted that four pulses arrive at the pulse widener input (V3) for each pulse-pair entering the second delay line, DL2.

The pulse widener (V3A and V3B) output pulse is coupled to the clipper amplifier, V4, which clips both the top and bottom portions of the pulse. The output pulse of this stage is transferred to a pulse amplifier, V5.

The output of the pulse amplifier, V5, is applied along dual paths to shunt-regulated amplifiers comprising V6 through V8 and V10 and V11, respectively. The circuits are arranged to develop a large output voltage with good output regulation, despite variations in load impedance. The signal fed to the discharge gate has a duration of 7 μs and an amplitude of 450 volts.

A 12-kv potential is applied to the series connection of the charge diode, V13, and discharge gate, V14. The gate pulse from the shunt-regulated amplifier (V6 through V8) allows the discharge gate stage to operate for 7 μs. During this time a capacitor, C31 (1 μf at 13,500 VDC) is allowed to discharge. The 100-volt, 7-μs output pulse is passed to the klystron, V15, for purposes described later. At the end of the gate pulse, the capacitor, C31, quickly recharges to its normal value.

FREQUENCY MULTIPLIER-OSCILLATOR

A portion of the second delay line, DL2, in figure 14-11 output is fed through cathode followers (V2A and V2B) to a pulse widener circuit (V7A and V7B) in the frequency multiplier-oscillator chassis. For each pulse entering the amplifier-modulator chassis from the coder-indicator, two pulses are fed from the delay line to enter the pulse widening circuit of the frequency multiplier-oscillator. These pulses are approximately 1.5 μs wide at half amplitude. One pulse is delayed from the other by an amount determined by the taps on the delay line. The delay can be varied from 1.8 to 2.7 μs in steps of 0.15 μs. The nominal value is 2.3 μs.

The operation of the V7A and V7B pulse widener circuit is essentially the same as that of the pulse widener circuit, V3A and V3B, discussed earlier. The leading edge of the first pulse from the delay line, DL2, initiates the leading edge of the pulse widener output. The leading edge of the second pulse from the delay line triggers the pulse widener circuit to generate the trailing edge of the circuit output pulse.

The positive pulse of 2.3 μs duration from the pulse widener circuit (V7A and V7B) is

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applied to a clipper amplifier, V8. This stage provides amplification and a means of adjusting the pulse baseline. The output signals from the clipper amplifier are fed simultaneously to two pulse amplifiers. One of the pulse amplifiers is located in the amplifier modulator chassis and the other in the frequency multiplier oscillator chassis.

The output of the pulse amplifier in the frequency multiplier oscillator is fed to a low-pass shaper network, which has a characteristic impedance of approximately 550 ohms. When a pulse of 2.3 μs is coupled to the network, a symmetrical output pulse of suitable duration (3.5 μs) and amplitude to trigger the shunt-regulated amplifier, V10 and V11, is produced.

The shunt-regulated amplifier is provided to ensure a constant d-c pulse level to the shaper output as applied to the final doubler, V6, and likewise to ensure constant pulse amplitudes. The signal applied to the input of the final doubler is 3.5 μs wide, and of approximately 200 volts in amplitude. The effects of this signal on the final doubler will be discussed later.

The second output of the clipper amplifier, V8 (in the frequency multiplier-oscillator chassis) is fed through a pulse amplifier in the amplifier modulator to a second shunt-regulated amplifier. This circuit is driven by the outputs of two different pulse wideners (V3 and V7, through respective clipper amplifier and pulse amplifier circuits). The V3 output has already been described, a pulse of 7 μs duration. The V7 output, which is 2.3 μs wide, is obtained from the pulse widener circuit of the frequency multiplier-oscillator. As a result of the overlap, a pulse wider than either input is developed by the combination of the two pulses. The resultant pulse, 8.5 μs in width, is used for keying the first doubler, and for blanking the receiver during the interval that a pulse is being transmitted. Thus, the possibility of feedback from the transmitter to the receiver is reduced.

The crystal oscillator, V1, an over-stabilized, plug-in, unit, and operates in the range from 40 to 51 mc, depending on the particular channel. A portion of the oscillator output is fed through frequency multipliers to a "balanced tee" mixer in the receiver as the local oscillator signal. This signal is multiplied 24 times through the chain of frequency multipliers, as discussed earlier.

Another portion of the oscillator output is fed through a buffer amplifier, V2, to the first pulse doubler, V3. The latter stage is keyed by the 8.5-μs pulse from the shunt-regulated amplifier comprising V10 and V11. Keying is provided at this point to diminish the physical size of the following r-f amplifiers, and of the power supplies, by reducing the duty cycle in the r-f stages.

The output of the first doubler, V3, is further multiplied in a tripler, V4, and a doubler, V5, and fed to a final doubler stage, V5. Thus, the crystal oscillator frequency is multiplied 24 times.

The final doubler is keyed by the 3.5-μs pulse from the V10 and V11 shunt-regulated amplifier in the frequency multiplier oscillator chassis. It should be recalled that one of these pulses appears for each pulse delivered from the coder-indicator.

A directional coupler (not shown) is used at the input to the final doubler to facilitate tuning. A directional coupler is a device used to sample the power in the incident and/or reflected waves of an r-f transmission system. For low standing-wave ratios, the power of the incident wave is a rough indication of the power delivered to the load.

The klystron, V15 (fig. 14-13) is a special high gain amplifier. It operates as a three-cavity amplifier, and depends on the changes introduced in the velocity of the stream of electrons by alternately slowing some of the electrons and accelerating others, and using the transit time between the two points to produce an alternating current. Cavity tuning is accomplished by mechanically changing the gap spacing.

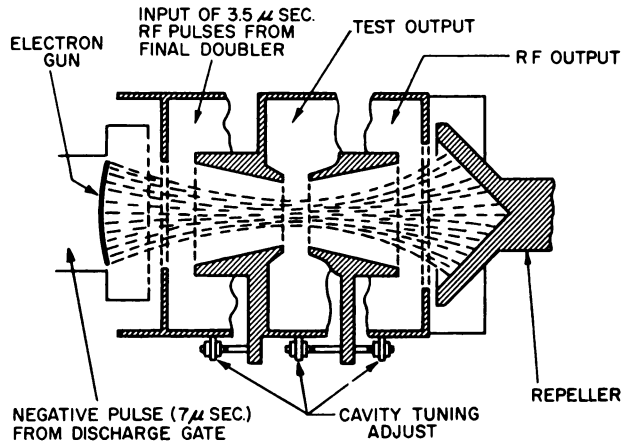

Figure 14-13.—Sectional diagram of klystron tube.

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The electrons emitted from the concave cathode of the klystron tend to converge at a focus point. However, by mutual electron repulsion in the cavity space, the electrons are caused to spread out, as shown in the diagram. The beam comes down to a minimum and again spreads so that it does not have a uniform diameter throughout its length. The over-all power gain of the klystron is slightly more than 100 at its operating range of approximately 1000 mc.

It is well to recall at this point that the output of the coder-indicator (in one second) consists of 3600 pulse-pairs, with 12-μs spacing between pulses of a pair. Thus, the pulses fed through the first delay line, DL1, and pulse clipper, V1 (fig. 14-11) are paired. For each pulse entering the second delay line, DL2, four pulses will leave the circuit. Two of these are fed to circuits in the amplifier-modulator, while the other two are fed to the frequency multiplier-oscillator. The pulses are spaced at intervals dependent upon the taps on the line. For a pulse-pair from the coder-indicator, eight pulses are generated, each appearing at a different time.

The two pulses passing through the frequency multiplier-oscillator, which originate from a single pulse input to the line, trigger the pulse widener circuit (V7A and V7B) to produce the wider (2.3 μs) pulse. The second pulse of the coded pair causes a similar action to generate another 2.3-μs pulse. The leading edges of the widened pulses have the same spacing as the leading edges of the coded pair. These pulsers are again widened as described and subsequently used to key the first pulse doubler, V3, and final doubler, V6.

Also recall that the clipper amplifier, V8, output (2.3- μs pulses) are fed to the shaper network (fig. 14-14) through a pulse amplifier. These pulses are delayed 1.3 μs and widened to 3.5 μs by the shaper circuit. The 3.5-μs pulses are fed through the V10 and V11 shunt-regulated amplifier of the frequency multiplier oscillator to key the final doubler, V6 (fig. 14-11).

Finally, the remaining two pulses from the second delay line (for a single-pulse input) are used to produce the 7 μs output of the amplifier modulator. This output, passing through the clipper amplifier, V4, pulse amplifier, V5, and shunt-regulated amplifier (comprising V6 through V8), is applied from the discharge gate, V14 to beam (key) the klystron, V15.

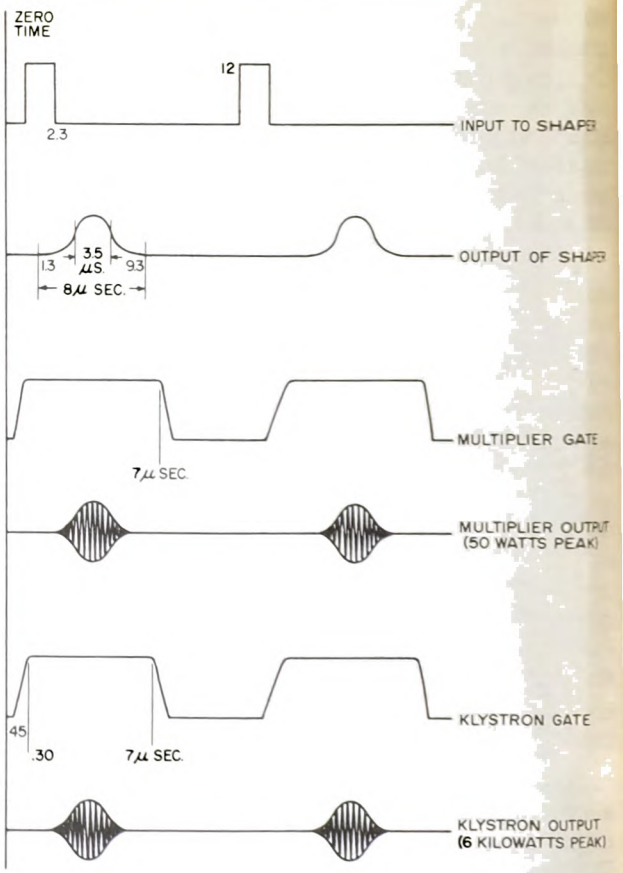

Figure 14-14.—Transmitter pulse sequence.

The 1.3-μs delay in the shaper output is necessary so that the 3.5-μs modulating pulse will arrive at approximately the center-time of the klystron beam pulse (fig. 14-14). This prevents undesirable distortion of the output waveshape.

The klystron (fig. 14-11) output (approximately 6-kw peak) is coupled to the duplexer. The duplexer is essentially a tuned filter used as a protective device to allow the single antenna to receive and transmit. Directional couplers (not shown) are placed in the input and output lines to monitor the power of the incident and reflected waves. From the duplexer, the pulsed output signals are applied to the antenna.

ANTENNA SPEED CONTROL

As discussed previously, the speed of the rotating cylinders carrying the parasitic reflectors must be kept at a constant speed of

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)0 rpm. This is true since the speed of the ıtenna determines the modulating frequencies. ɔ control the speed, a tachometer and a speed ɔntrol system are employed with both shore-ısed and shipboard radio beacons. The tach-neter shows an output of 810 cps when the ıtenna cylinders are spinning at the required)0 rpm.

The speed control system (fig. 14-15) con-sts essentially of a frequency sensitive de-ce, a preamplifier with one saturable reactor, ıd a power amplifier with three saturable ɜactors. Two error-sensing step-down trans-ɔrmers are employed to couple and impedance .atch the tachometer output to the 805- and 15-cps frequency discriminators. The fre-ıency discriminators sense the error in the ıchometer frequency from the desired 810 cps, ıd produce two currents whose difference is roportional to the error frequency. The fre-uency discriminators are actually two series esonant circuits, one tuned to 805 cps, while ıe other is tuned to 815 cps. These circuits re pretuned, and hermetically sealed, and annot be readjusted in the field.

Separate step-up output transformers are ısed to couple the frequency discriminator utputs to their respective bridge rectifiers. The variation in the direct current from the ridge rectifiers of the high and low resonant ircuits as the tachometer frequency is changed s illustrated in figure 14-16. Note that the two urves, A and B, are displaced equally on ither side of the nominal tachometer frequency. The curve at C represents the difference be-ween the low and high frequency circuits. This lifference indicates the error curve, which ndirectly controls the speed of rotation of the ıntenna.

Notice that the difference in the two curves f figure 14-16 is zero at 810 cps. Any small leviation from this frequency produces an error urrent, which has the proper direction and hase to return the speed of the antenna cyl-nders to 900 rpm, ± one-half percent.

The stabilizing networks function (fig. 14-15) n conjunction with the preamplifier saturable eactor to form a phasing network. The circuit peration is such that the difference of the nmfs in the preamplifier produces the error utput. A stabilizing inductor is used to elimi-ıate excessive overshoot of the antenna cylinder peed.

The preamplifier functions as a multiwound saturable reactor whose output alternating current is rectified to provide a positive feedback, and to provide the d-c control current for the three 3-phase saturable reactors in the power amplifier.

Single-phase input power for the antenna speed control circuits is supplied from the line, and is applied from a power transformer to the bias current bridge rectifier.

The preamplifier output is applied through a control and bias current rectifier to three separate output saturable reactors. Each of the reactors is in series with one phase of the 3-phase power line. Thus, the line current supplied to the speed control (spin) motor is directly controlled by the effective impedance of these reactors. For low values of d-c control current to the spin motor, the saturation is low and the impedance is high. The motor voltage is decreased and the motor will slow down. Conversely, for large values of control current the motor voltage will be increased, due to the lowered reactor impedance, and the motor will speed up.

A bias rectifier is placed across one phase of the input power to supply the three separate bias windings. Each bias winding is shunted by a single dry disc rectifier (not shown). The rectifier damps oscillatory currents and suppresses harmonics of the line frequency that are induced in the windings of the reactor.

Three power factor correction circuits in a wye connection are provided to correct the power factor of the spin motor. Ordinarily, only capacitors would be required, but since the amount of capacity necessary would make the equipment too large, an autotransformer is provided. This autotransformer multiplies the capacitance by the turns ratio squared, thus increasing the effective amount of capacitance in the correction circuit.

ANTENNA SERVO SYSTEM

The bearing servo system (fig. 14-17) is used with the stabilized shipboard antenna. The angular position signals for true north are continuously transmitted from the ship's gyro compass through the ship's bus to the radio beacon. The magnetic variation subassembly, located in the coder-indicator unit, is manually adjusted to correct the ship's gyro compass information for magnetic north. The signals maintain the radio beacon's 15-cps reference

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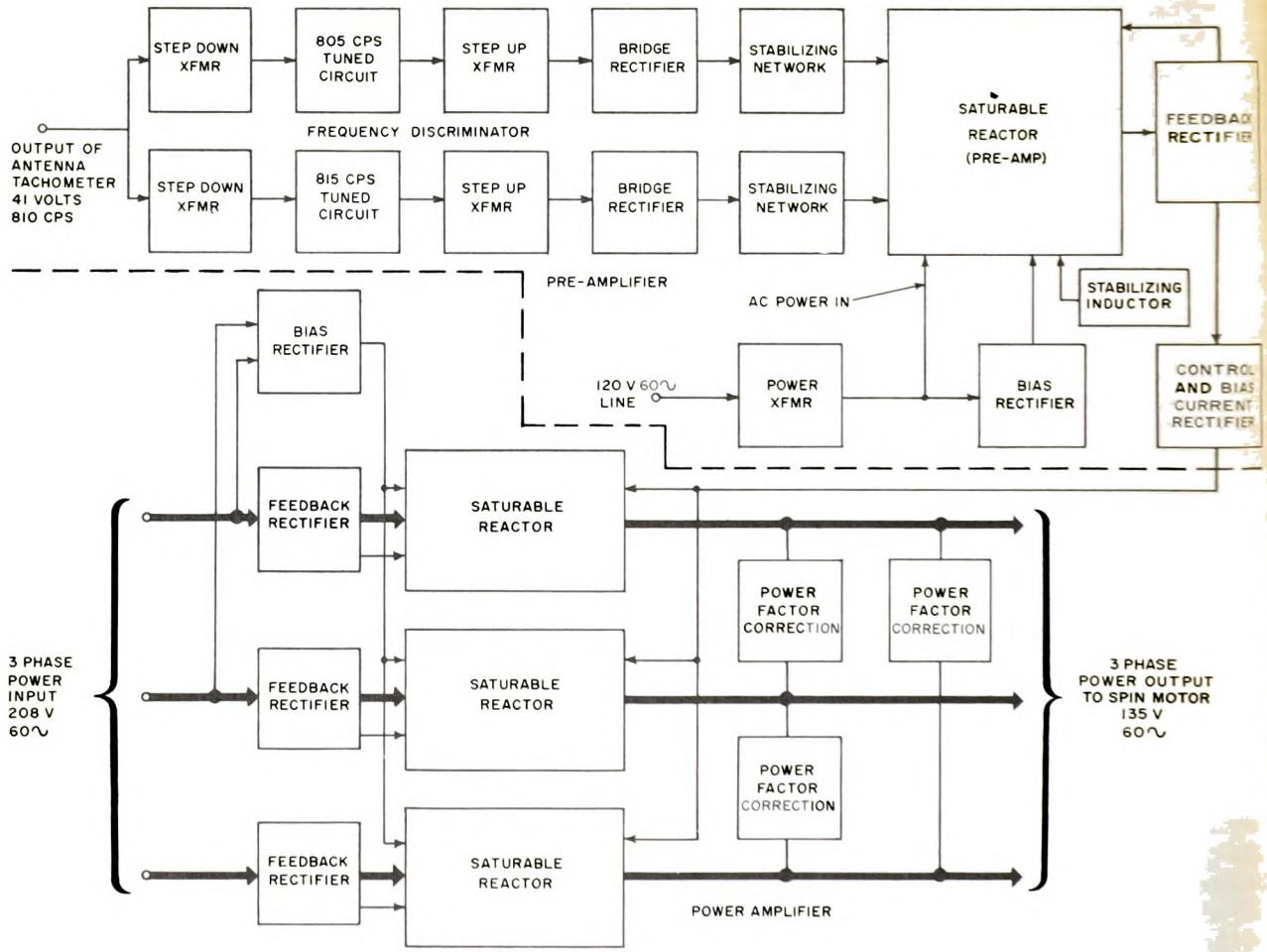

Figure 14-15. —Antenna speed control circuits, block diagram.

pulse coil subassembly fixed with respect to corrected (magnetic) north.

The magnetic correction signals are manually set into 1- and 36-speed differential generators of the magnetic variation assembly. The sum of the gyro compass and magnetic variation signals is forwarded to the azimuth 1- and 36-speed control transformers in the antenna base.

As long as the rotors of the control transformers are at the same effective angular position as that of the differential generators, after summing the signals of the ship's gyro compass and the magnetic variation signals, no control transformer output is sent to the bearing amplifier. If the equilibrium is upset due to displacement of the rotors of the differential generators to new positions by changing the setting of the magnetic variation unit, or should the ship's course change, an error voltage will be induced and transmitted from the control transformer to the bearing servo amplifier.

The bearing servo amplifier, having received its signal, sends power to the servo motor, and causes it to rotate in the proper direction to restore the rotors of the control transformers to the zero signal position. At the same time the control transformers are being restored to equilibrium, the gear train rotates the pickoff coils of the pulser and plate assembly enabling the 15-cps reference of the transmitted signal to be oriented to magnetic north as dictated by the ship's course and the magnetic variation subassembly.

The magnetic variation subassembly is mounted on the front panel of the coder-indicator for the convenience of the operator.

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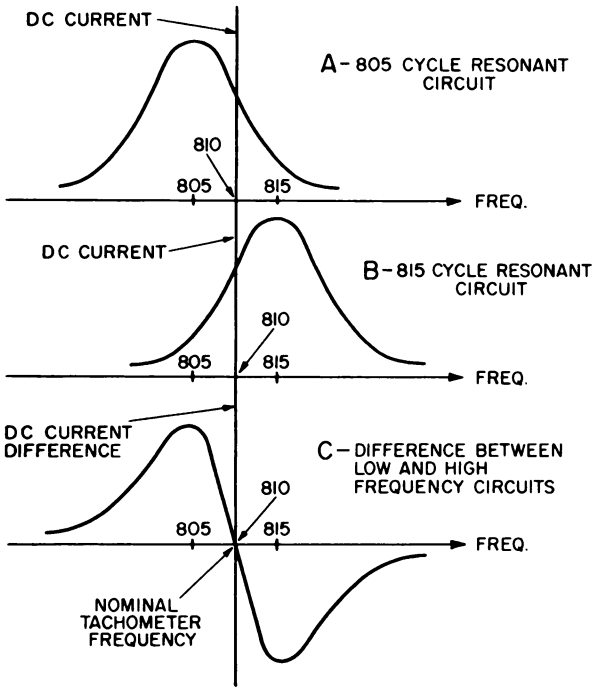

Figure 14-16.—Antenna speed control circuits, analysis of waveshapes.

It is composed of two synchro-differential generators, the rotors of which are geared together in a ratio of 36 to 1; a dial calibrated in 360° of bearing, which is connected to the rotor of the 1-speed differential generator; and a dial calibrated from 0 to 10° attached to the rotor of the 36-speed synchro. Both are displayed concentrically on the front panel of the coder-indicator.

Rotation of the dial of the magnetic variation subassembly by the operator results in a command to the servo amplifier, which in turn causes the pickoff coils of the pulser and plate assemblies on the antenna to rotate a corresponding amount. This action fixes the 15-cps reference of the transmitted signal to the new magnetic north position indicated.

SYNCHRO ZEROING

The error voltage of the servo loop is sensed by the control transformer and sent to the servo amplifiers (fig. 14-18). To ensure that the servo loop always tries to orient itself to the position indicated by the input signal, the circuit has been designed to have only one stable point on the synchro voltage curve.

The error voltage versus degrees of error for the 1- and 36-speed synchros is shown in figure 14-19. The zero point on the curve is the desired point of orientation, or stable point, of the servo loop. A stable point may be defined as one where the voltage curve crosses the zero axis and changes from a negative to a positive phase with increasing angle.

Two stable crossover points (0° and 180°) occur in 360°, with the 1- and 36-speed signals crossing the zero point simultaneously. In figure 14-19, A, if the system is between 175 and 180° (where correspondence is 0°) when a signal is introduced, the 36-speed CT signal (below the axis) will cause the servo to run in one direction (assume ccw) and drive the system to 180° where it will lock in to correspondence at a false zero.

To eliminate this condition and to ensure a single point where the system will lock in (0°) a low-amplitude 60-cycle voltage (called anti-stickoff voltage) is introduced. This action moves the cross over point from 180 to 185° (fig. 14-19, B).

If the system is between 180 and 185° when the signal is introduced, the 36-speed CT signal (above the axis) will cause the servo to run clockwise (opposite to ccw because the phase of the signal is reversed) and drive the system back through 180° toward 0°. If the system is slightly above 185° when the signal is introduced, the 36-speed CT signal will cause the servo to run in a ccw direction to drive the system to 360°, which is equivalent to the true zero point of correspondence.

The 185° crossover point is now unstable in the combined 1- and 36-speed voltages because it has the wrong phase and slope (positive feedback) required to meet the conditions of a stable point (negative feedback) as described above. Therefore, the servo loop will always try to orient itself at only one point (0°) on the voltage curve, the point represented by the angular displacement of the ship's compass and the magnetic variation subassembly (for shipboard radio beacons).

BEARING MAGNETIC AMPLIFIER

Either of two different designs of bearing magnetic amplifiers may be furnished with the equipment. A block diagram of the unit designed by Magnetic Amplifiers, Inc., is shown in figure 14-18. The block diagram of the second type, designed by the Federal Telephone and Radio

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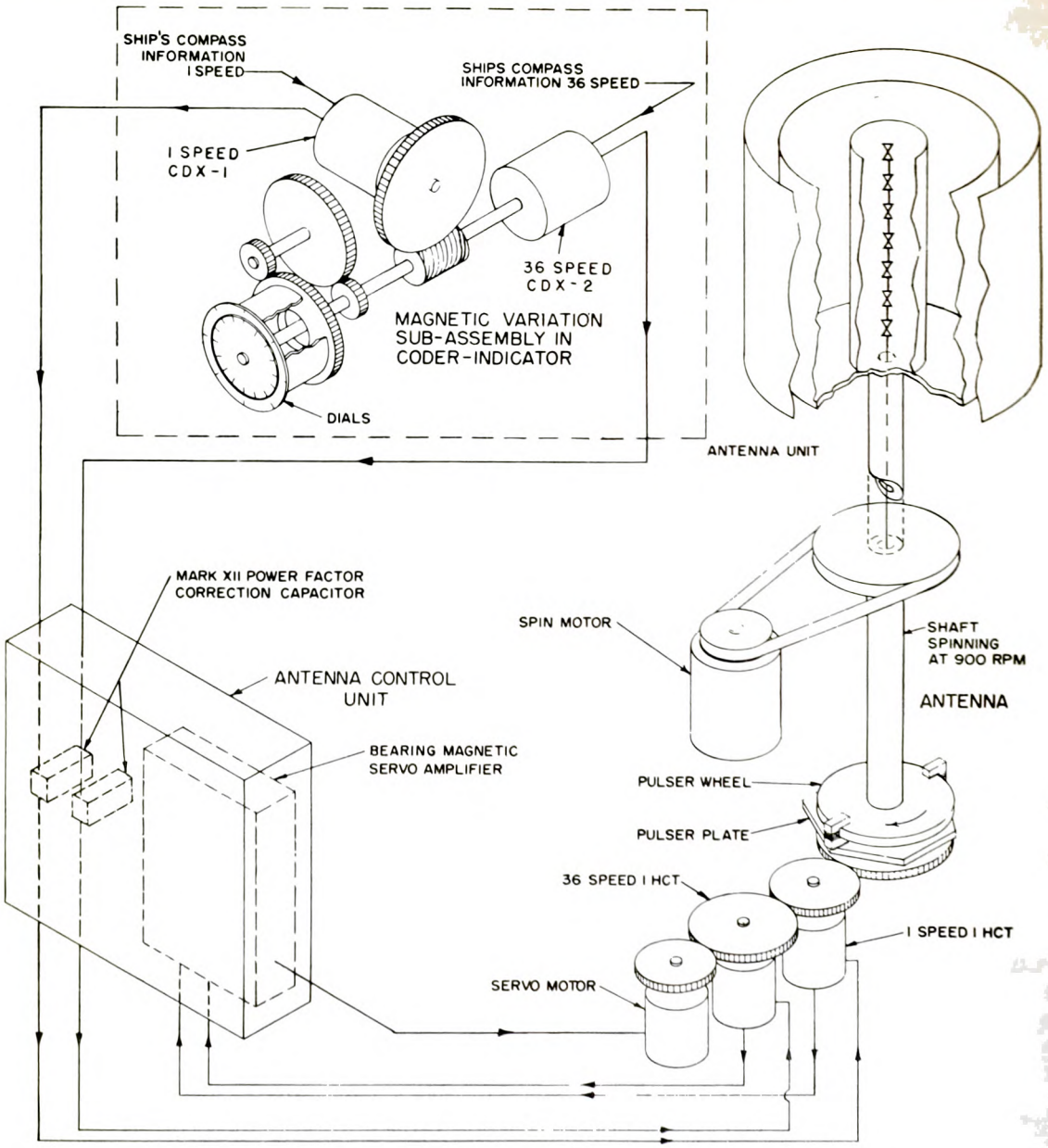

Figure 14-17.—Bearing servo system, functional block diagram.

Company is shown in figure 14-20. Although the over-all purpose of the separate units is the same, the construction of the two is sufficiently different to warrant separate discussion. The magnetic amplifier type is discussed first.

In order to detect the reversal in phase of an a-c signal, it is necessary to compare it to a fixed reference. The phase-sensitive detector produces an output voltage across a load that is proportional to the magnitude of the difference between the reference and error voltage. As long as the polarity of the error voltage remains the same, the direction of the current through the load will also remain the same. If the polarity of the error voltage reverses with respect to the reference voltage, then so does the direction of the current through the load. Thus, the polarity of the a-c load voltage will reverse with a reversal in the phase of the a-c error. The phase-sensitive detector will also

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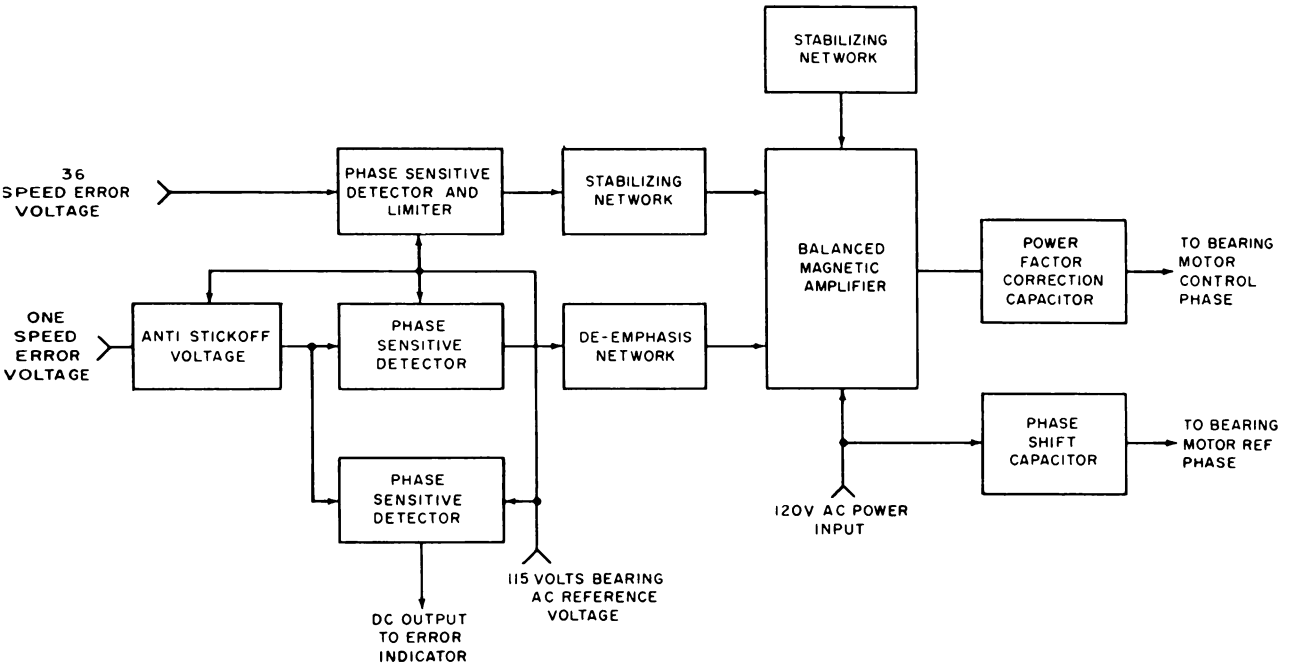

Figure 14-18.—Bearing amplifier (Magnetic Amplifiers, Inc. Design) block diagram.

function as a limiter if the amplitude of the error voltage equals or exceeds the reference voltage.

One phase-sensitive detector is employed for the 1-speed synchro, and one for the 36-speed error signal. A third detector (fig. 14-18) has its input paralleled with the 1-speed detector, and is used to supply direct current to the bearing (azimuth) error-indicator meter on the front panel of the coder-indicator.

The 1-speed error voltage is modified by the addition of the antistickoff voltage previously described. This signal is added to the 1-speed voltage by means of an isolating transformer (not shown).

The deemphasis network in the 1-speed circuit is used to obtain a nonlinear resistance. At low input levels, the resistance of this network is high, and for high levels, the resistance is low. Thus, with low error inputs the gain of the 1-speed circuit is decreased, and, conversely, when the error is large the 1-speed circuit has a larger control since the lowered resistance permits a much larger controlling current. It is necessary to deemphasize the 1-speed signal voltage at small errors since it is less accurate than the 36-speed signal and would introduce additional error into the system.

A stabilizing network provides an advance in the phase of the input current to compensate for the inductive characteristics of the servo system. The circuit is commonly used with servos for the purpose of increasing stability.

The balanced magnetic amplifier combines the 1- and 36-speed voltages and supplies its output (along dual paths) to the power factor correction circuit and a phase-shift capacitor, respectively. A stabilizing inductor comprises a phase-lag network for one of the windings of the amplifier.

The portion of the magnetic amplifier output fed to the power factor correction circuit passes through a 3.75-microfarad capacitor (not shown). This capacitor neutralizes the inductance of the magnetic amplifier and the motor winding. The motor control voltage is thus approximately in phase with the line voltage.

A second phase-shift capacitor is employed to shift (in quadrature) the phase of the reference voltage. This capacitor also has a value of 3.75 microfarads, and functions to shift the phase as required to provide starting torque for the bearing motor reference phase.

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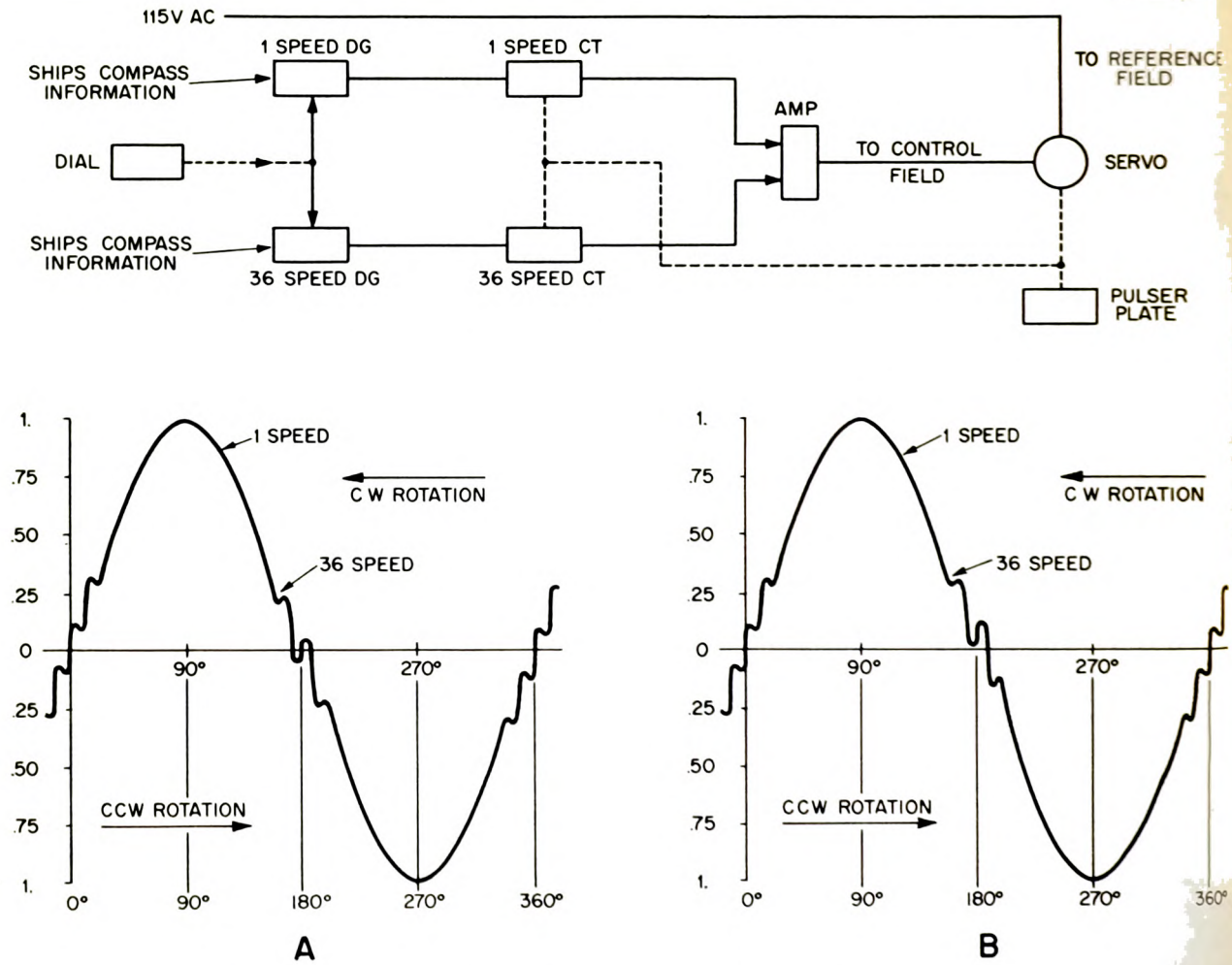

Figure 14-19.—Error voltage versus degrees of error, graphic analysis.

The bearing magnetic amplifier (fig. 14-20) Federal Telephone and Radio Company, provides power for the control phase of the bearing motor in a slightly different manner than that just discussed. The 1-speed error signal, with added antistickoff voltage, is amplified by the switching preamplifier. The d-c output from this amplifier, proportional to the error voltage, is used to block the switching amplifier when the 1-speed error signal is greater than 2.5°. A relay, operated by the output from the two amplifiers when the 1-speed error is 2.5° or less, switches the input to the phase-sensitive detector from the 1-speed input to the 36-speed input.

The phase-sensitive detector is essentially the same as that described earlier, except that the error and reference voltages are interchanged. If the incoming signal is in phase with the reference, the output to the push-pull preamplifiers will be increased. The respective output signals from the two amplifiers are 180° out of phase, and are fed to the push-pull power amplifiers.

No phase shift occurs in either the preamplifiers or the power amplifiers. As the output of the top preamplifier decreases, the output of the top power amplifier will also decrease. Similarly, the output of the lower power amplifier increases. When the two power amplifiers are thus thrown out of balance, the output phase of the bottom amplifier predominates, and is fed into the control phase of the servo motor.

Degenerative stabilization is obtained by taking the a-c output voltage drop across the power amplifier windings, rectifying this voltage,

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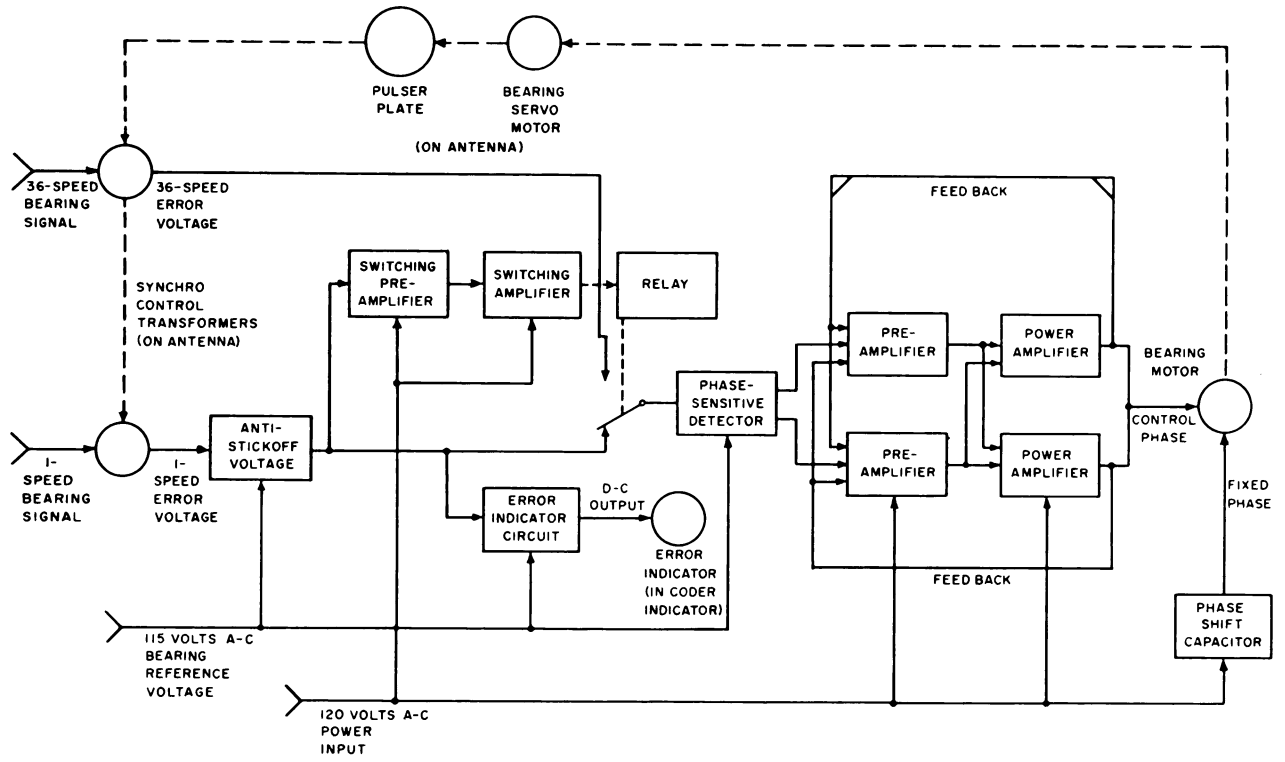

Figure 14-20.—Block diagram of bearing magnetic amplifier (Federal Telephone and Radio Company).

and feeding it back to the preamplifiers in a direction that will oppose the preamplifier with the higher output, and aid the preamplifier with the smaller output.

The bearing amplifier includes a phase-shift capacitor used to alter the phase of the fixed (reference) voltage applied to the bearing motor. This action makes the fixed phase in quadrature with the control phase to produce bearing motor starting torque when the control phase is energized. Also included is a circuit that operates an error indicator located on the coder-indicator unit. The circuit employs a second phase detector that receives the error signal from the 1-speed voltage before the switching relay. A built-in limiter circuit is provided to protect the meter.

ROLL AND PITCH SERVO SYSTEM

GENERAL DESCRIPTION

Roll and pitch information is fed through the ship's bus to the antenna stabilization servo system by their respective control transformers (fig. 14-21). The error signal developed across the control transformer is sent to the magnetic amplifier. The magnetic amplifier, in turn, supplies power to the servo motor for a specific direction of rotation, as determined by the error signal.

The servo motor, through its associated train, moves the antenna to its corrected position. At the same time, the gear train nulls the error signal at the control transformer through the 1:2 gear box train. A rate feedback loop is provided through the tachometer to eliminate oscillation.

MAGNETIC AMPLIFIERS

The magnetic amplifiers for the roll and pitch are identical electrically, and function similarly to the bearing magnetic amplifier. Roll and pitch amplifiers of two different designs are furnished with this equipment. The block diagram of one type, manufactured by Magnetic Amplifiers, Inc., is shown in figure 14-22. The block diagram of the other type, manufactured by the Federal

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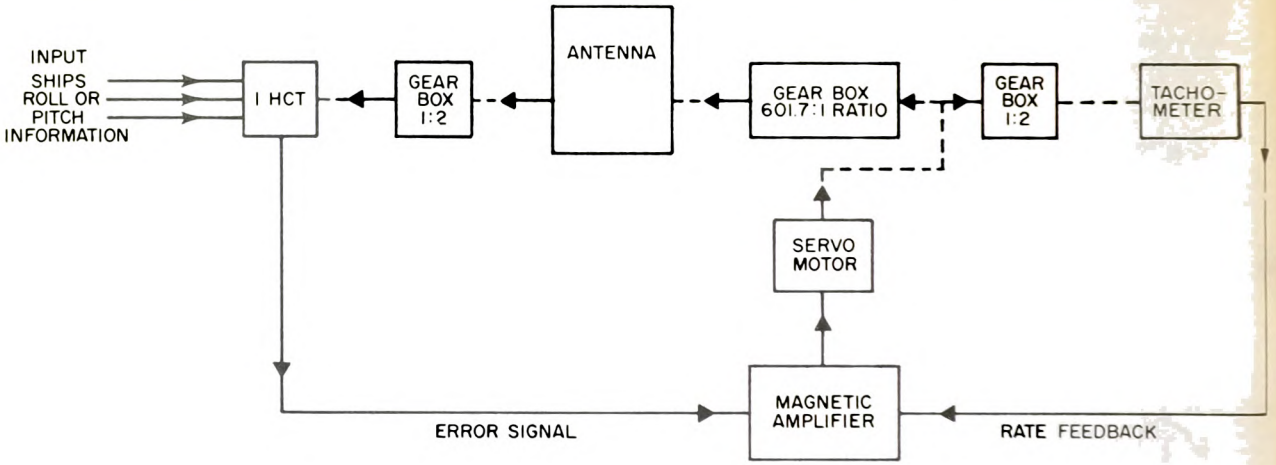

Figure 14-21.—Simplified roll and pitch magnetic amplifier block diagram.

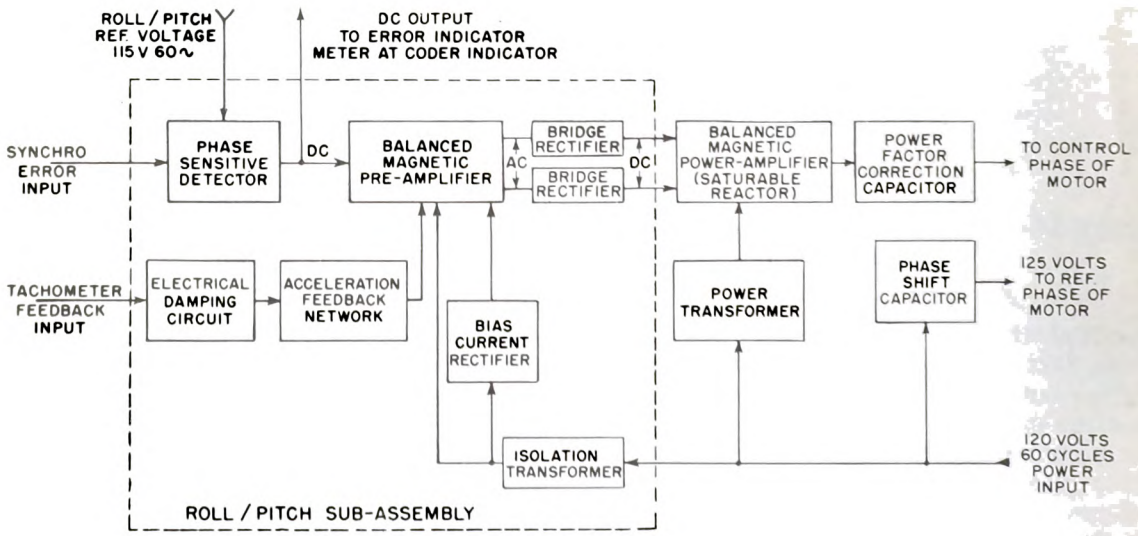

Figure 14-22.—Roll or pitch (Magnetic Amplifier, Inc., Design) block diagram.

Telephone and Radio Company, is shown in figure 14-24.

Separate motor-driven tachometers (fig. 14-23) are employed in the roll and pitch circuits. The speed of the tachometers depends on the roll and pitch and is also controlled by the correction signal from the servo motor (fig. 14-24).

The synchro error voltage, which is obtained from the ship's gyro compass via the roll and pitch control transformers in the antenna base, (fig. 14-23) is fed into a phase-sensitive detector (fig. 14-22) of the type already discussed. The d-c output is the control current for the balanced magnetic preamplifier.

The error and tachometer feedback signals are applied (as separate inputs) to the balanced magnetic preamplifier. The tachometer normally supplies an output of approximately 4.5 volts per 100 rpm. Any tendency of an increased speed of the antenna (as a result of roll or pitch) causes a high feedback voltage to be developed and the tachometer output to be increased. This action lowers the antenna speed. A slower antenna speed causes a smaller feedback voltage to be developed, and the magnetic amplifier

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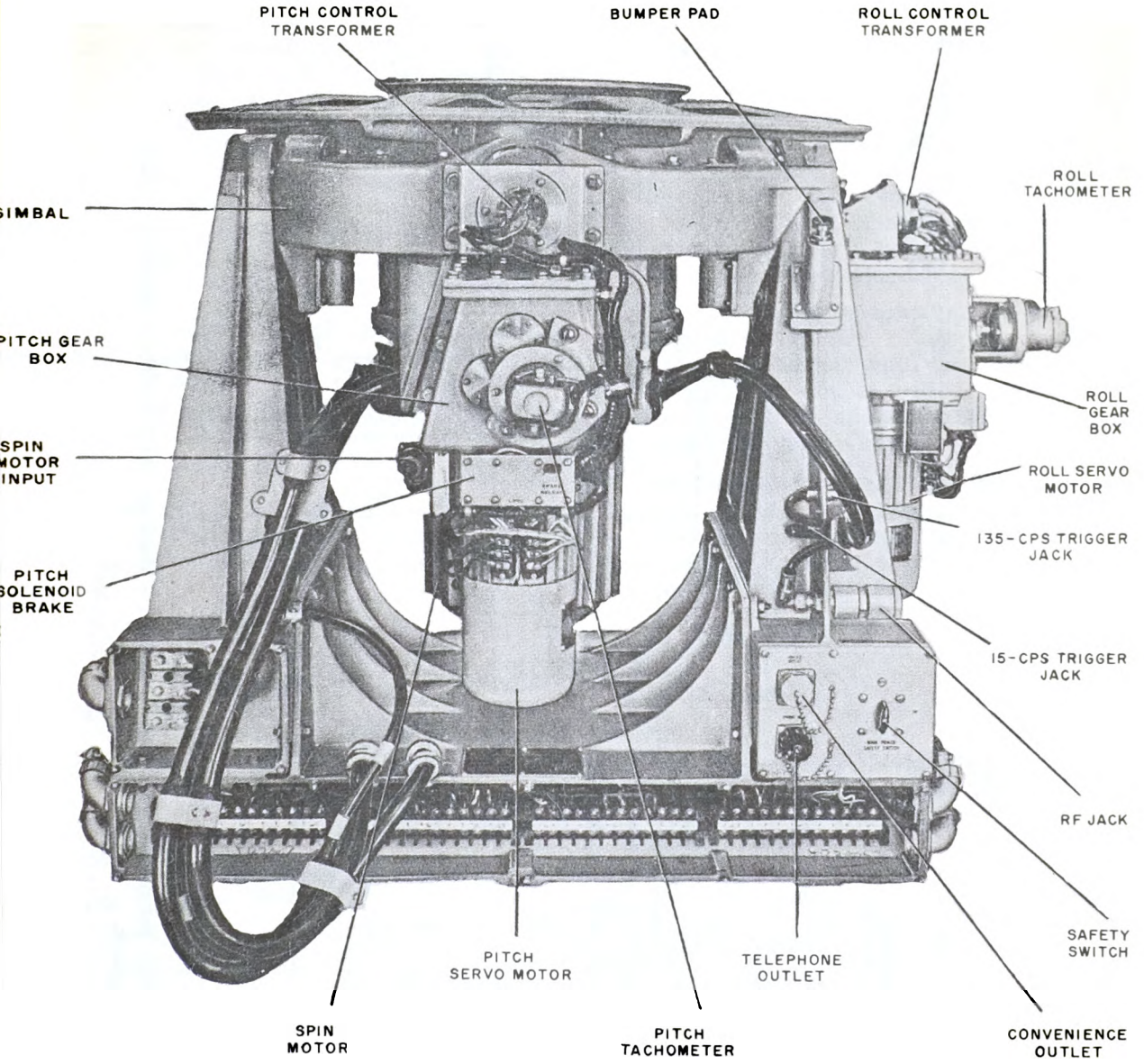

Figure 14-23.—Antenna base showing roll and pitch components.

output raises the speed of the antenna back to its normal value. The tachometer (degenerative) feedback voltage counteracts the magnetic amplifier circuit in a manner that prevents overshoot or undesired oscillations in the loop.

The error and tachometer feedback circuits (fig. 14-24) are isolated. The electrical damping circuit controls the amplitude of the tachometer feedback input. The isolation impedance in the tachometer circuit also functions to change the tachometer voltage from rate feedback to acceleration feedback, thus preventing overshoot, and increasing the stability of the servo system.

A bias current rectifier (fig. 14-22) is used to provide a d-c bias for the magnetic amplifier. The output of the magnetic amplifier is rectified in two paralleled bridge rectifiers, which provide positive feedback, and control the bias current for the next stage.

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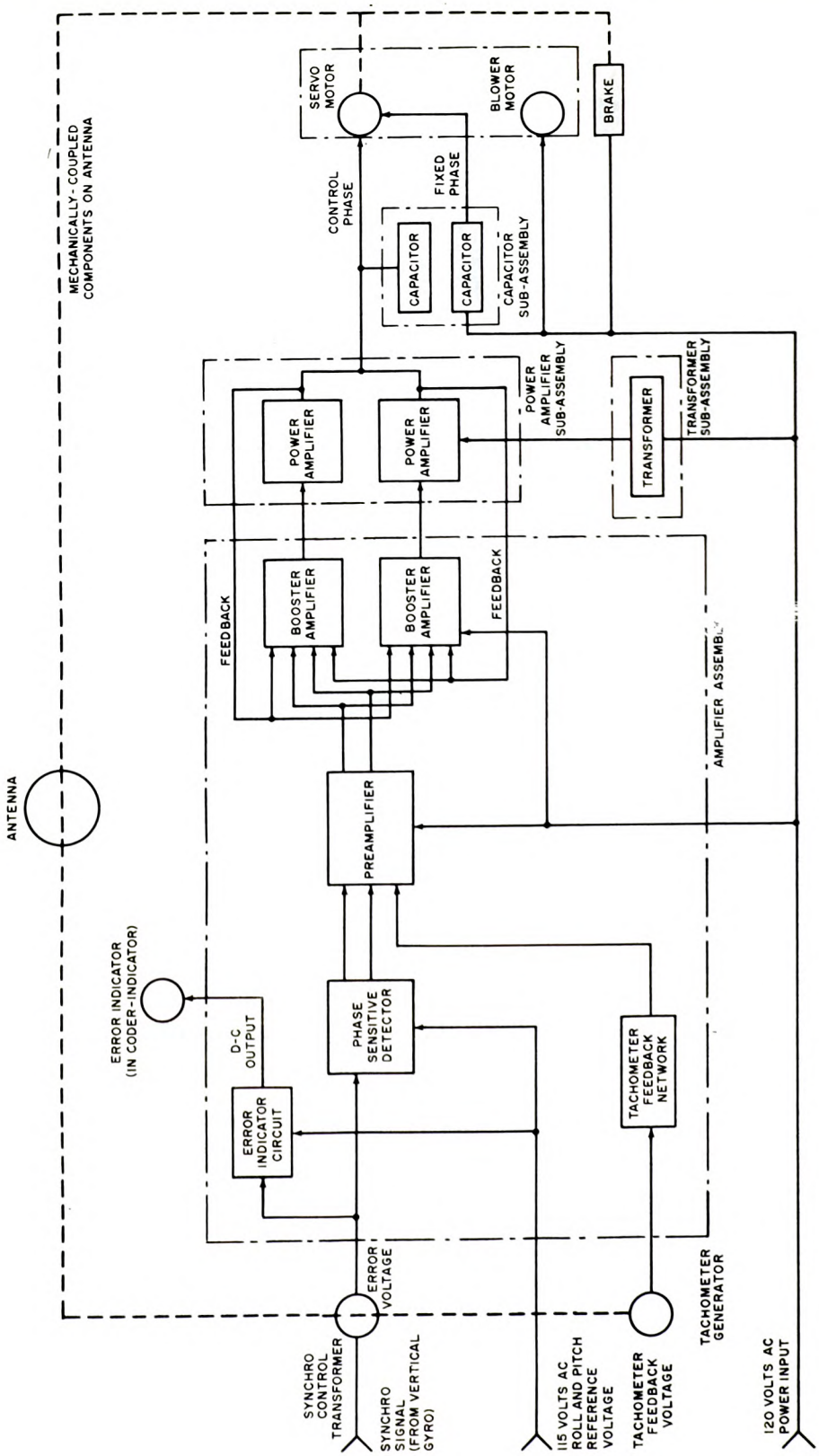

Figure 14-24.—Roll or pitch magnetic amplifier (Federal Telephone And Radio Company Design) block diagram.

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The balanced magnetic power amplifier (saturable reactor), along with feedback rectifiers and power transformer, comprises the output circuit that feeds the roll and pitch motors. A power factor correction capacitor is placed in series with the line to the control phase of the motor to shift the phase relationship of the current-voltage output to approximate that of the line. Also, a phase-splitting capacitor is placed in series with the line voltage that shifts the reference (fixed) phase of the servo motor voltage 90° to facilitate starting.

The two designs of roll and pitch magnetic amplifiers made by the different manufacturers as described above are not sufficiently different to warrant separate block diagram discussion. An observation of the block diagram (fig. 14-24) will show the similarity.

The roll and pitch amplifiers of this design are identical. The components of each amplifier are divided between four subassemblies as indicated.

POWER SUPPLIES

LOW-VOLTAGE POWER SUPPLY

The low-voltage power supply (fig. 14-25) contains three full-wave rectifiers. The rectifier plate transformers are connected to the 120-volt, single-phase power through the LVPS plate contactor. The contactor is operated by the control circuit (discussed later). A red transformer indicator lamp is lighted when the rectifier plate transformers are energized.

The B+ protector functions to prevent the application of the B+ voltage from the low-voltage power supply until the -375- and -400-volt supplies are in operation. This allows the bias voltage to be applied to the tubes prior to the positive B supply voltage.

Conventional capacitor input filters are connected in the output circuits of the rectifiers. A gas-filled voltage regulator tube is used (in each source) to apply a constant d-c potential to the control grid of an electronic voltage

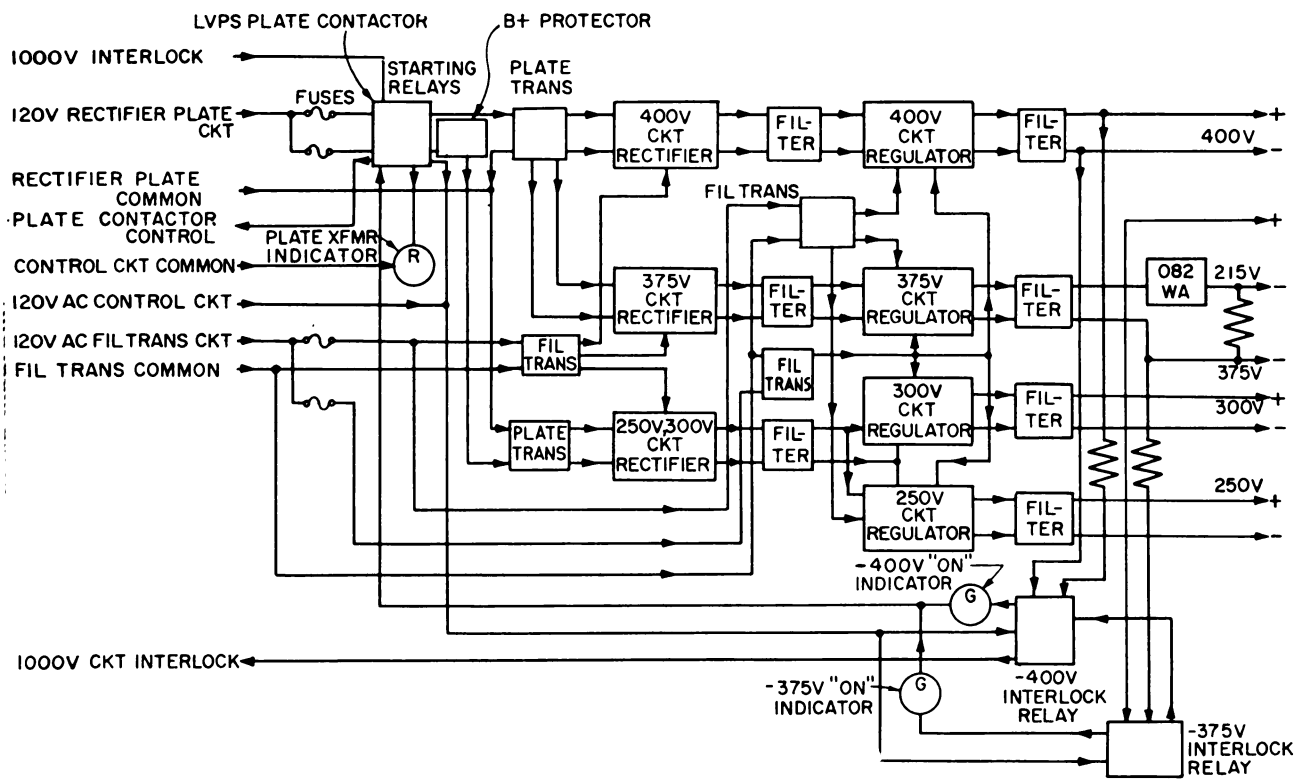

Figure 14-25.— Low voltage power supply, block diagram.

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regulator circuit. The four regulators (one for each source) are essentially the same, differing only in the value of the resistors and in the number of series regulator tubes connected in parallel. Capacitor filters are connected in the output circuits of the regulators.

Gas-filled regulator tubes are also connected in series with resistors across the -375-volt source. A tap on this circuit provides a regulated -215-volt output.

Two magnetic interlock relays are used, one across the -375-volt output, and the other across the -400-volt supply. One set of contacts on each relay is connected in series with part of the control circuit to prevent the medium voltage power supply from functioning unless the -375- and -400-volt outputs are available. Another set of contacts on each relay closes the circuits of two green indicator lamps when the -375- and -400-volt outputs, respectively, are present.

MEDIUM-VOLTAGE POWER SUPPLY

The medium-voltage power supply (fig. 14-26) contains one full-wave rectifier employing two tubes. The rectifier plate transformer is connected to the incoming 120-volt, single-phase power through the starting relay (plate contactor). The contactor is operated by the control circuit as will be described later.

The rectified output is filtered and applied to an electronically controlled voltage regulator circuit. Five gas-filled voltage regulator tubes (not shown) are used to apply a constant cathode potential to a control tube, which senses and corrects any error in the 1,000-volt output amplitude. A single capacitor filters the 1,000-volt regulated output.

A separate regulator circuit is connected across the 1,000-volt divider to provide a 500-volt regulated output. A positive 700 volts nominal is obtained by voltage dropping across two series resistors before the output connection is made to the power supply terminal board. The voltage across the load connected between the 850-volt terminal and ground may vary, while the voltage from the regulator side of the two series resistors to ground will be maintained at 1,000 volts.

The medium voltage power supply contains an overload relay, which functions to remove the plate voltage when an excessive current is sustained. A lag or time delay action in this relay is necessary so that large charging currents that occur when the equipment is started will not activate the overload relay.

When the solenoid of the supervisory relay (connected across the 1,000-volt regulated output) is energized, one set of contacts closes part of the control circuit to apply 300 volts to the screen grids of the multiplier tubes in the

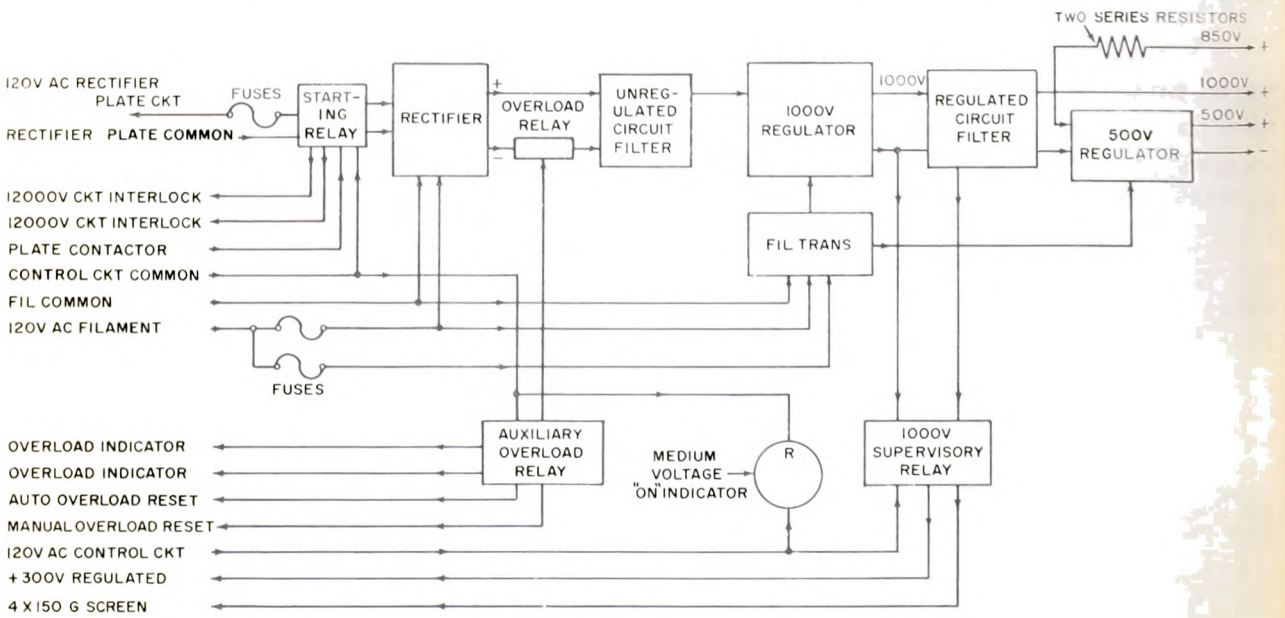

Figure 14-26.—Medium-voltage power supply, block diagram.

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radio beacon transmitter (see figure 14-11). Another set of contacts completes the circuit to turn on the red "medium voltage ON" indicator located on the front panel of the medium voltage power supply.

HIGH-VOLTAGE POWER SUPPLY

The high voltage power supply (fig. 14-27) contains a 3-phase rectifier, which receives a 9,220-volt plate power via a plate transformer mounted separately in the power supply test set cabinet (see figures 14-1 and 14-2). The rectifier supplies a 12,000-volt output, which is filtered by two 1.0μf capacitors (not shown). The high-voltage power supply does not employ a voltage regulator.

The supervisory relay is connected across a portion of the bleeder circuit, and is thus energized by the 12,000-volt output. A variable resistor (not shown) in series with the relay solenoid makes it possible to adjust the solenoid voltage to a suitable value. The contacts of the relay close a portion of the control circuit to allow application of screen voltage to the discharge gate tube in the radio beacon only after the high voltage has reached approximately 10 kv.

Two elapsed time indicators are mounted on the front panel of the high-voltage power supply unit. One begins operation when the plate contactor is closed and power is supplied to the unit, and thus records the total time that the 12,000-volt direct-current has been supplied to the klystron. A red indicating light is connected in parallel with the elapsed time indicator and warns the operator that the 12,000-volt output is present.

The second elapsed time indicator records the total time that the high-voltage power supply filament transformer is energized to supply the klystron. It also indicates the time that power is supplied to the other filaments of the transmitter.

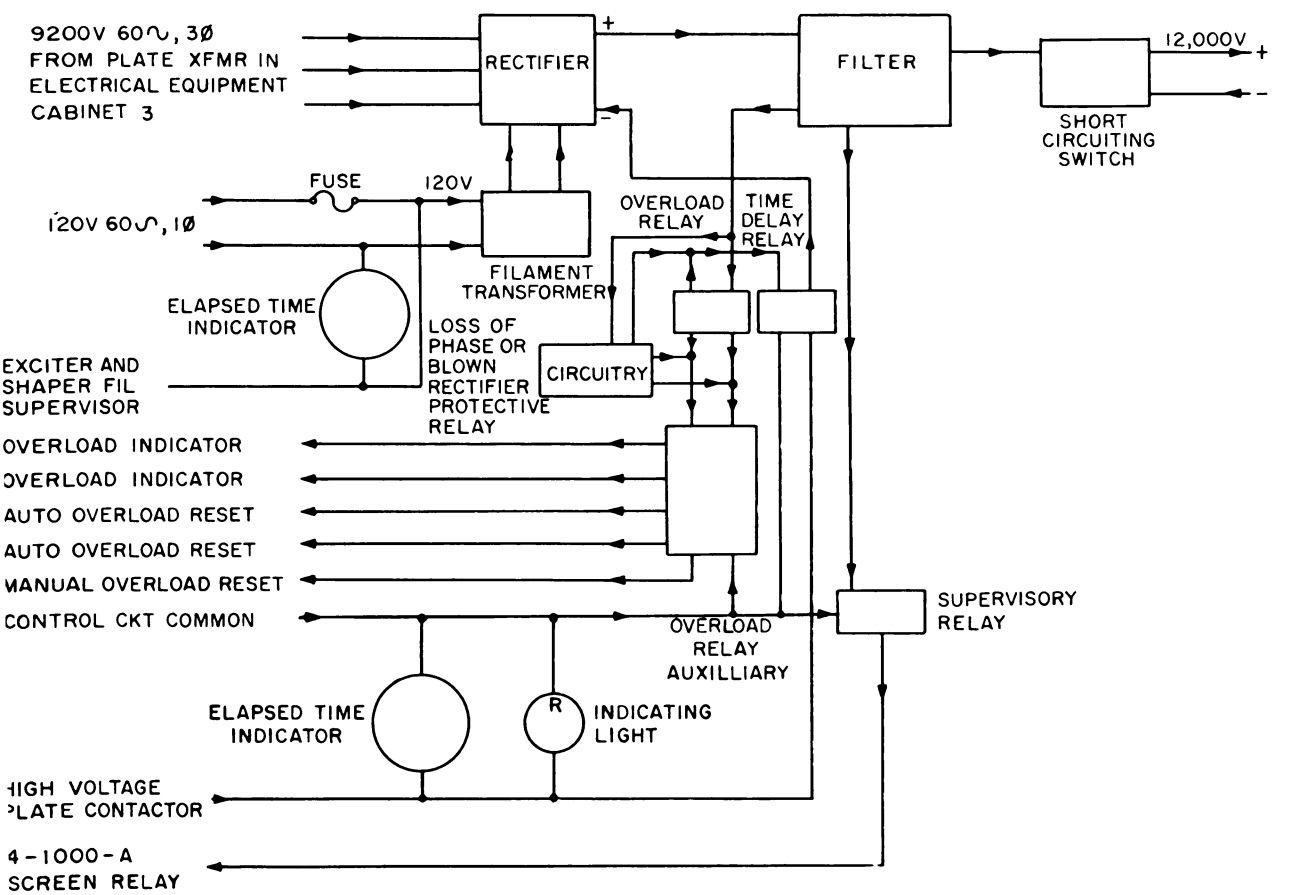

Figure 14-27.—High-voltage power supply block diagram.

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A protective (shorting circuit) switch is provided on the high-voltage power supply drawers that shorts, through a resistor, the 12,000-volt output and discharges the filter capacitors when the drawer is pulled from the cabinet.

Protection against loss of phase or power or a bad rectifier tube is provided automatically by the protective relay. The time delay relay functions to keep the protective circuits de-energized until starting transients have decayed.

The overload relay is connected across a portion of the rectifier plate circuit. Excessive currents energize the relay to remove the 12,000-volt output.

POWER DISTRIBUTION AND CONTROL CIRCUITS

The power distribution and control circuits (fig. 14-28) function in such a way that the various components of the equipment are energized in their proper sequence, and that the failure of one component will result in the de-energizing of other components that might be damaged by its failure. Safety interlocks on component units and compartments in which high voltages are present disconnect parts of the system when the units are pulled out or the compartments opened.

The main power supplied to the equipment is 208-volt, 60-cycle, 3-phase, and 120-volt (or 117-volt) 60-cycle single phase. Land-based equipments may obtain this power from a commercial source. The shipboard equipment uses a delta-wye stepdown transformer connected to the 440-volt, 3-phase, 60-cycle ship's power supply. Rectifiers in the various components convert the a-c to d-c, as discussed. An additional 117-volt, single-phase, 60-cycle power supply may be used for convenience outlets and for the crystal oven.

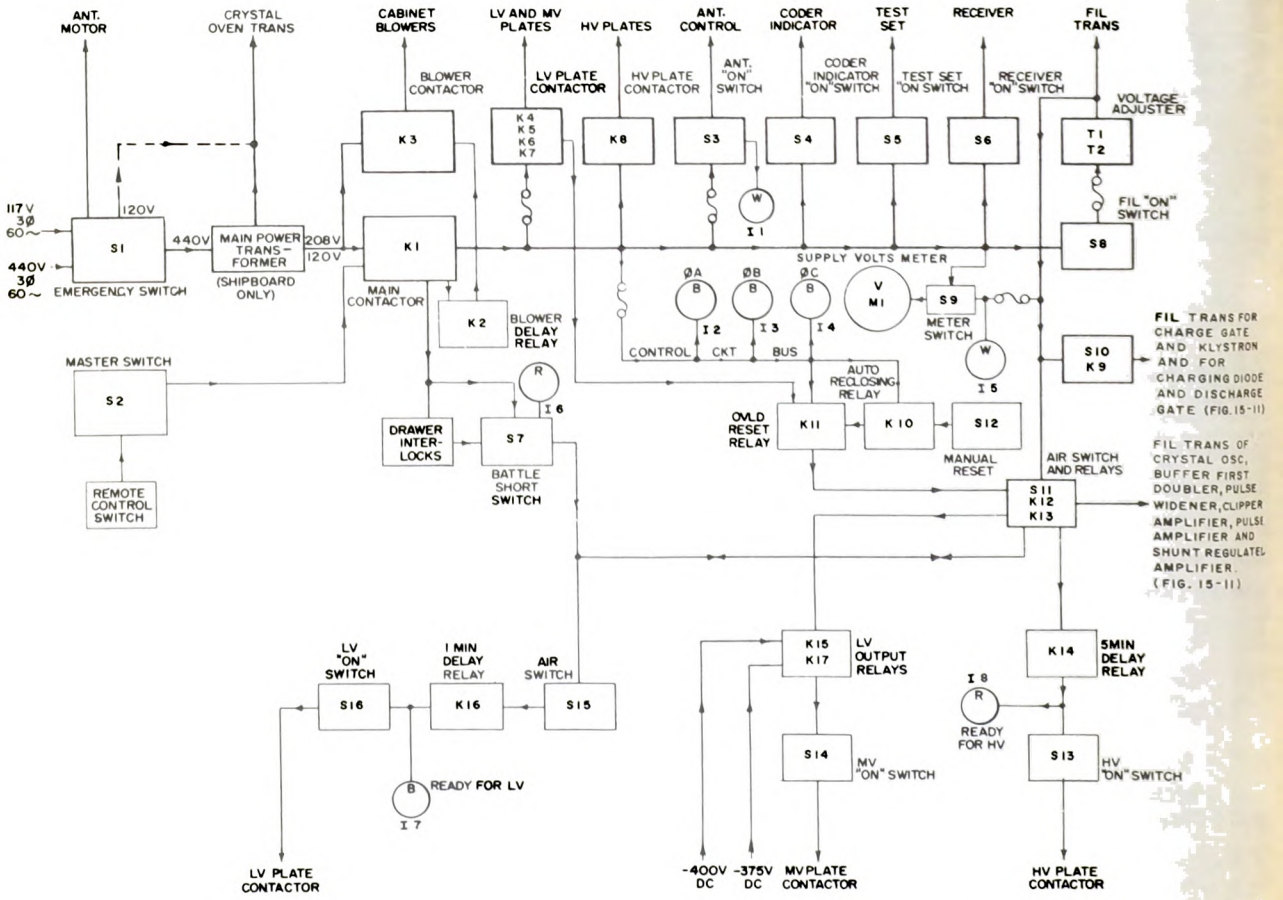

Figure 14-28.—Radio beacon power distribution and control circuits, block diagram.

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BLOCK DIAGRAM

The functioning of the power supply and control circuits is as follows: When the operator closes the emergency switch, S1 (mounted on the receiver-transmitter cabinet), power is connected to the antenna motor contactor. The 208-volt, 3-phase, 60-cycle power is connected via the main power transformer) to the terminals of the main contactor, K1, and the blower contactor, K3.

120-volt, single-phase power is connected to the crystal oven transformer, which regulates the temperature on the crystal in the crystal oscillator (fig. 14-11). This power may be obtained from the convenience outlet circuit, or in shipboard applications) from the secondary of the main stepdown transformer.

If the master switch, S2 (fig. 14-28), is now operated to the STANDBY position, main contactor K1 will close, connecting 208-volt, 3-phase, 60-cycle power to the terminals of the high-voltage plate contactor, K8. Then, 120-volt, single-phase, 60-cycle power is connected to the "Fil-On" switch, S8, the low- and medium-voltage plate contactors, K4, K5, K6, and K7, the control circuit bus, the antenna control switch, S3, the test set switch, S5, the coder-indicator switch S4, the radio receiver switch, S6, the "Main Power ϕA ϕB ϕC" indicator lights I2, I3, and I4, respectively, and the coil of the blower contactors, K3 via K2.

The blower contactor closes, connecting the frequency multiplier-oscillator blower, the power supply test set cabinet blower, and the receiver-transmitter cabinet blower to the 208-volt, 3-phase power. When the power is shut off by the master switch, the blower contactor is kept closed by the blower time delay relay, K2, for one minute after the main contactor, K1, has opened so that the equipment will not be damaged by excessive heat, and to dissipate rapidly the heat stored in the hottest parts of the equipment. The equipment must not be shut down by the emergency switch until one minute has elapsed after returning the master switch to the OFF position.

With these circuits operating, the operator may now close the following switches:

1. The antenna-on switch, S3, energizing the antenna control unit. If the spin motor switch on the antenna control is closed, the motor will be activated, causing the antenna to rotate.

2. The coder-indicator switch, S4, may be closed to energize this unit.

3. The test-set-on switch, S5, which energizes the associated test circuits.

4. Receiver-on switch, S6, may be closed to energize the radio receiver circuit.

5. The Fil-on switch, S8, may be closed, energizing the filament transformer and the white "fil" indicating light, I5. The operator may adjust the filament transformer voltage to exactly 120 volts by means of the adjust-for-120-v knob on the control panel, which operates the voltage adjusting transformers, T1 and T2. Ordinarily, the supply volts voltmeter, M1, is connected to the adjusted filament transformer voltage by the meter switch, S9. The meter may also be used to check the unregulated voltage by connecting it to the line through the use of the meter switch.

When the Fil-on switch is closed, the filament transformers of the low, medium, and high voltage power supplies, the filament transformers in the frequency multiplier-oscillator, and the transformers in the amplifier-modulator are immediately energized. Filament transformers and white indicating lights in the frequency multiplier-oscillator are energized through K12 (in the control duplexer) if the air switch in the frequency multiplier-oscillator is closed by the action of the blower.

Relay K12 closes K13, which, in turn, energizes the time delay relay, K16, which prevents the low and medium voltage power supply plate voltages from being applied until the filaments have been connected for one minute. The low-voltage power supply activates the medium-voltage power supply, as will be discussed later.

K9 energizes two filament transformers and a white indicating light in the amplifier-modulator if the air switch, S10, is closed by the action of the blower in this unit. Simultaneously, time delay relay K14 is energized, providing a 5-minute delay before the high voltage power supply plates (and indirectly the klystron) can be energized.

The operator may leave the master switch at STANDBY indefinitely, and the equipment will remain in the condition just described. If the master switch is now moved to the ON position, the unit interlocks are connected to the 120-volt, a-c control power. The low-voltage power supply will be ready to be turned on if:

1. The units and compartment covers are in place, or the battle-short switch, S7, is closed (which is done only in extreme emergencies) or the individual interlocks have been manually

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closed as may be the case during tuning procedures.

2. The air switch, S15, has been closed by the action of the power supply test set blower.

3. The one-minute time delay relay, K16, is closed.

4. The blue light ready-for-LV is lit.

The operator might have advanced the master switch to ON without stopping at STANDBY. In which case the equipment will arrive at this condition after the various time delays have been completed.

If the LV ON switch, S16, is now closed, the low voltage plate contactor will be energized, causing the low-voltage power supply to function. The medium-voltage power supply will also be energized at this time if the following conditions exist:

1. K15 is energized by the -375-volts, d-c output of the low-voltage power supply.

2. Relay K17 is energized by the -400-volt, d-c output of the low-voltage power supply.

3. The medium voltage power supply overload relay (fig. 14-26) is deenergized.

4. Overload reset relay K11 is deenergized.

5. K12 is energized through the air switch, S11, by the action of the blower.

6. The MV ON switch, S14, has been turned on by the operator.

The d-c output of the medium voltage power supply closes the screen supervisory relay (fig. 14-26), which connects +300 volts d-c, obtained from the low-voltage power supply, to the screen of the tubes in the frequency multiplier-oscillator.

The high-voltage power supply plate contactor will close, connecting 208 volts, 3-phase power to the primary of the plate transformer if the following conditions exist:

1. The starting relay (fig. 14-26) of the medium voltage power supply is energized. The starting relay completes the interlock circuit for the 12,000-volt supply.

2. K12 is energized through the air switch, S11, by the action of the frequency multiplier-oscillator blower.

3. Overload reset relay, K11, is deenergized.

4. The contacts of the 5-minute time delay relay, K14, are closed. This relay is energized at the same time as the filament transformer, which heats the filaments of the klystron and charge gate tubes, and also the transformer, which heats the filaments of the discharge gate and the charging diode. (See figure 14-11.)

The blue indicating light, I8, shows ready-for-HV, which signifies that the conditions above have been met.

5. The overload relay in the high voltage power supply (fig. 14-27) is deenergized.

The HV ON switch, S13, has been turned on by the operator.

The high voltage d-c output energizes the supervisory relay in the high voltage power supply unit. This relay, in turn, energizes the discharge gate screen relay located in the amplifier-modulator unit with power obtained from the 120-volt, a-c control circuit bus. The screen relay connects +700 volts (nominal) d-c, obtained from the medium voltage power supply, to the screen of the discharge gate tube. Depending on the load conditions, this voltage may vary between 800 and 900 volts.

The above discussion describes the normal functioning of the control and power distribution circuits. Particular emphasis on the following features may be helpful:

1. All the ON switches on the front panel may be closed before the master switch is advanced to the STANDBY or ON position.

2. If the master switch is moved to ON (all ON switches being closed), the equipment will be automatically energized in its proper sequence, a period of five minutes being required before it is fully energized.

3. Once the equipment has been fully energized (all ON switches remaining closed), the master switch may be moved back to STANDBY and then returned to ON, and the equipment will then be immediately fully energized. The equipment may therefore be left in STANDBY position when not required to send or receive information.

If the high-voltage power supply unit or the amplifier-modulator unit are pulled out, or the transformer and blower compartment panel is removed, the high voltage power supply will be disconnected while the low and medium voltage supplies will remain functioning. If the low and medium voltage power supply units, the radio receiver unit, and the frequency multiplier-oscillator unit are pulled out, the low, medium, and high voltage supplies will be disconnected.

Overload protection is provided for the medium and high voltage supplies. An overload relay in the medium voltage supply disconnects that unit and also the high voltage source. The overload relay in the high voltage supply disconnects that unit alone. An automatic reclosing relay, K10, restores the power through the

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esetting relay, K11, after three seconds. The utomatic relay will reset three times and then hut down until reset manually by the reset witch, S17. In either case, if the cause of the verload has not been corrected, the overload elay will trip again.

The interlocks-shorted switch, S7, (used nly in emergencies), shorts the drawer inter-ocks of the control circuit. S7 also shorts the nit interlocks on the radio receiver, and coder-ndicator, which disconnects the 120-volt a-c nput to their individual power supplies.

The individual unit interlocks (except for he high voltage power supply and the main mplifier-modulator interlocks) may be closed nanually with the unit pulled out when it is ecessary for tuning. Interlocks closed in this nanner are restored to normal functioning when he drawer is closed. Units in which high oltage is exposed have an additional contact n the interlocks which energizes a red warning ight associated with the particular circuit.

The equipment may be energized by a remote control switch (when provided) if all the ON switches on the front panel are closed and the master switch has been turned to the REMOTE CONTROL position. A white indicator lamp on the remote control switch panel shows that the equipment is ready for use. The remote control switch will have off, standby, and on positions, which will perform the same functions as the correspondingly marked positions of the master switch.

Two red indicating lights are provided on the remote control switch. One indicates that the low and medium voltage power supplies are functioning, while the other shows that the high voltage supply is operating.

The high voltage power supply contains a circuit that provides protection against loss of an input phase or loss of a high-voltage rectifier tube. A circuit is also provided to prevent spurious tripping of the protective circuits by transients. These protective functions are performed by the time-delay and loss-of-phase or blown-rectifier protective relays, and by a low-pass filter incorporated in the circuit.

QUIZ

1. Radio Set AN/URN-3 is used at shore activities or aboard ship to provide the radio beacon or ground portion of what type system?
2. What three major units comprise the AN/URN-3?
3. How many aircraft may simultaneously obtain navigational information in conjunction with a single installation of the AN/URN-3?
4. The frequency used for receiving in low band installations is how many megacycles above the frequency used for transmitting in the same channel?
5. The radio beacon output consists of what four types of information signals?
6. The time delay between the distance interrogation pulse-pair (transmitted from the aircraft) and the corresponding distance pulse-pair reply (transmitted from the radio beacon) is adjusted at the radio beacon to what value?
7. Bearing information is originated in what section of the radio beacon?
8. The pulse produced by the upper slug on the pulser plate (fig. 14-5, B) is used for what purpose?
9. The aircraft receiver utilizes the 15-cps and 135-cps modulation on the total received signal (fig. 14-9) in presenting what type of information?
10. The 135-cps amplitude modulation (fig. 14-9) is provided for what purpose?
11. Of all the signals transmitted by the radio beacon, what component alone is not generated within the beacon?
12. The squitter circuit (fig. 14-11) functions to produce a negative control voltage, the magnitude of which depends on what factor?
13. The north burst is triggered at a 15-cps rate by the trigger generated in what unit (fig. 14-5, B)?
14. The gate from the 135-cps gating multivibrator (fig. 14-11) permits the 135-cps burst gate to allow the passage of how many cycles of the oscillator frequency?
15. The output of the coder-indicator (fig. 14-11) during one second consists of how many pulse-pairs and with how much spacing between pulses of a pair?
16. The width of the pulse from the pulse widener circuit (in the amplifier-modulator fig. 14-11) is determined in what manner?
17. What is the nature of the shunt regulated amplifier outputs in the transmitter section (fig. 14-11)?
18. The crystal oscillator (fig. 14-11) is a plug-in unit, oven stabilized, and operates in what frequency range depending on the particular channel?

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19. What is the width of the pulse that keys the final doubler in the transmitter (fig. 14-11)?
20. The 1.3-μs delay in the shaper output (fig. 14-11) is necessary for what reason?
21. The klystron peak power output (fig. 14-11) is approximately what value?
22. The speed control system (fig. 14-15) consists essentially of what circuits?
23. What is the function of the autotransformer (fig. 14-15) used with the capacitors in the spin-motor power factor correction circuit?
24. The bearing servo amplifier (fig. 14-17), having received its signal, sends power to the servo motor, and causes it to rotate in the proper direction to restore the rotors of the control transformers to what position?
25. The one-speed error voltage (fig. 14-18) is modified by the addition of what other voltage?
26. The bearing magnetic amplifier (fig. 14-20) uses a phase-shifting capacitor for what purpose?
27. The B+ protector (fig. 14-25) functions to prevent the application of the B+ voltage from the low voltage power supply until what actions have taken place?
28. The d-c output of the medium voltage power supply closes the screen supervisory relay (fig. 14-26) which connects + 300 volts d-c from what supply to the screen of the tubes in the frequency multiplier-oscillator?

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CHAPTER 15

ELECTRONIC AIDS TO NAVIGATION: TACAN (PART II)

INTRODUCTION

As previously stated in chapter 14, Radio Set AN/URN-3 is used at shore activities or aboard ship to provide the radio beacon or ground portion of an air navigation system. The beacon transmits signals that the aircraft Radio Set AN/ARN-21 interprets into distance, bearing, and identification information. Thus, the pilot of the aircraft is able to determine his distance from the radio beacon, the bearing of the radio beacon from his aircraft (with respect to magnetic north), and from which beacon his transmission originated.

The theory of operation of the circuits in the receiver, coder-indicator, and transmitter of the Radio Beacon AN/URN-3 are considered in this chapter. A block diagram analysis of the entire AN/URN-3 equipment is given in Chapter 14, Electronic Aids to Navigation: Tacan (Part I) of this training course. It is assumed that you have read and understood the material contained in that chapter. Refer to the illustrations in chapter 14 when it is necessary to ascertain the position of the various circuits in the system. The receiver is shown in figure 15-1.

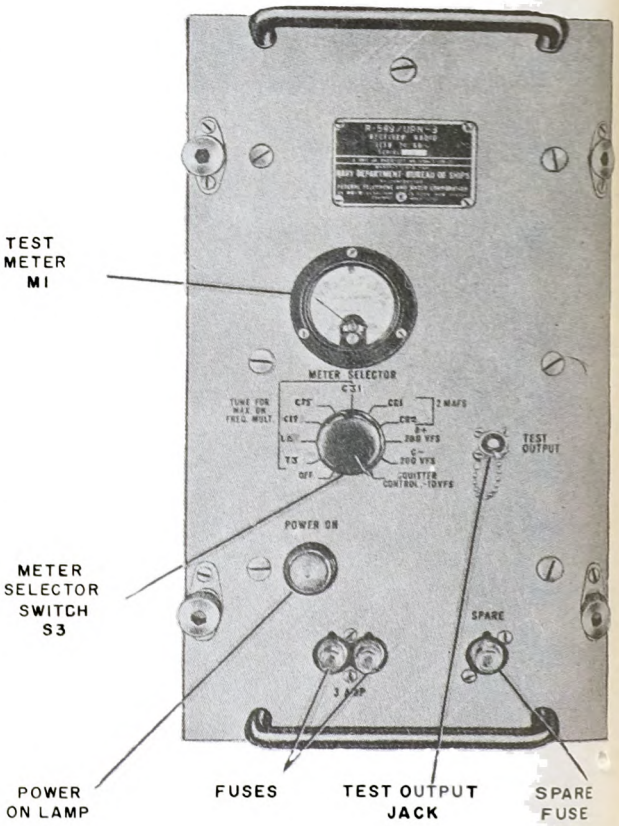

Figure 15-1.—Radio receiver, front panel view.

RADIO RECEIVER

INPUT CIRCUIT

Signals from the antenna are applied through a duplexer to a low-pass filter, Z1 (fig. 15-2). The duplexer is a filter network, which allows both the transmitter and receiver to be operated simultaneously from one antenna. The antenna and duplexer construction and operation are treated in detail later.

The received signals contain interrogating pulse-pairs transmitted by the aircraft. The decoding circuits of the receiver only respond to those pulse-pairs having the pulses of a pair, spaced 12 μs apart. For each interrogation pulse-pair decoded, the receiver sends a single pulse to the coder-indicator (fig. 14-11 of this training course).

MIXER

The interrogation pulse-pair signals from the low-pass filter are fed into one arm of a balanced magic TEE section, and applied to the two crystal diodes, CR1 and CR2, in phase. Because of the phase relationship of the interrogation pulse-pair signals at the output of the diodes, cancellation of this signal occurs across the output

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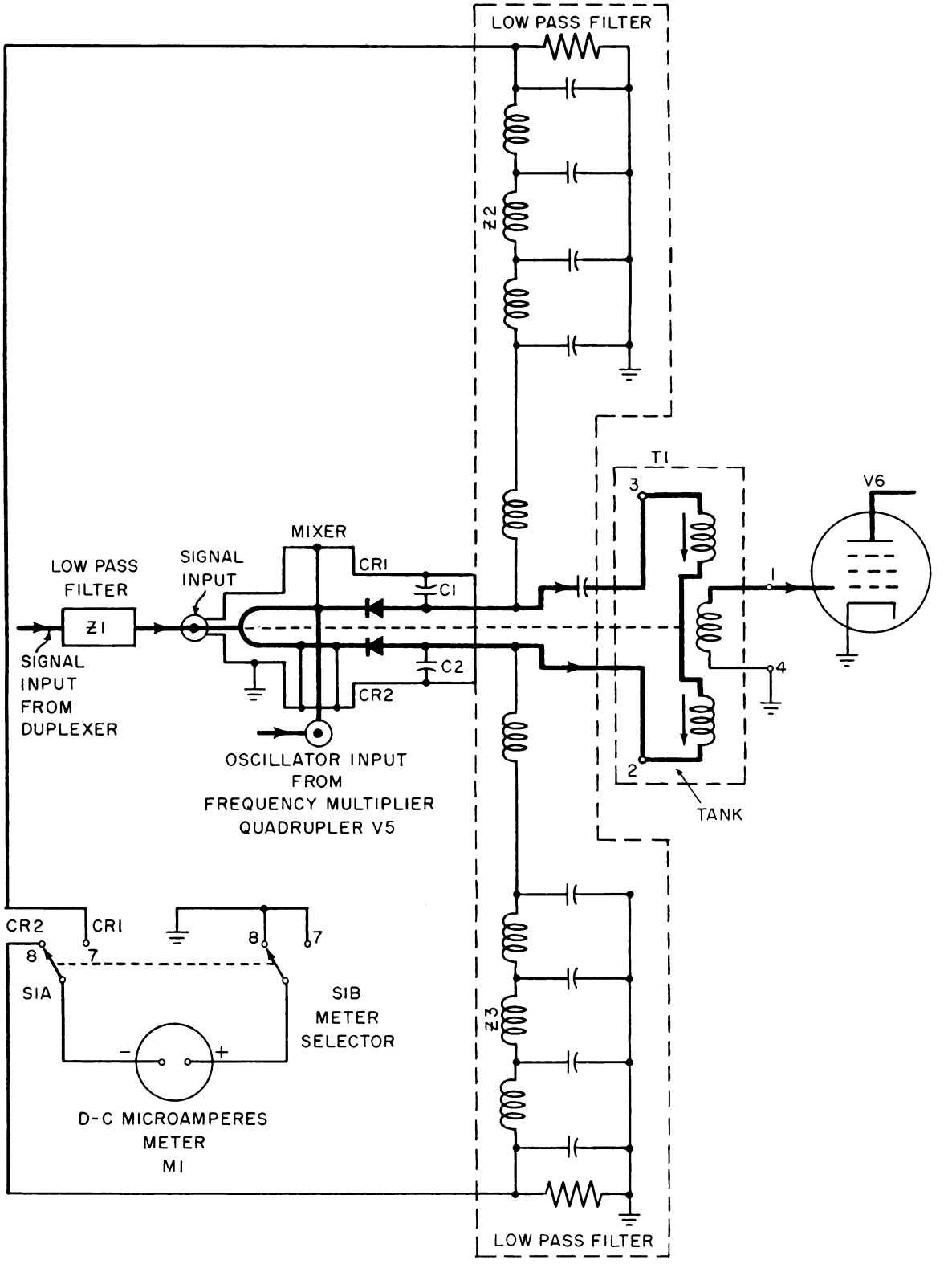

Figure 15-2.—Radio receiver balanced "TEE" mixer, schematic diagram.

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transformer. A local oscillator signal is applied to the cathode of one diode, which, in conjunction with the signal already present at this cathode, produces a rectified difference frequency across the mixer output transformer, T1.

The interrogation pulse-pair carrier frequency will be between 1025 and 1087 mc on the low band, and 1088 and 1150 mc for high-band reception. The local oscillator input to the mixer may be adjusted from 962 to 1213 mc. The heterodyning process at the mixer produces an intermediate (difference) frequency of 63 mc.

T1 is tuned by C1 and C2 to the intermediate frequency. The sum and original frequencies are shunted by these capacitors, allowing only the difference frequency of 63 mc to be developed by the tank comprising the T1 primary and the distributed capacitance of the T1 primary in shunt with C1 and C2.

A d-c microammeter, M1, may be switched to measure the CR1 or CR2 current by the meter selector switch, S1. Low pass filters Z2 and Z3 are connected in series with the input to the crystal current metering circuit to block the passage of unwanted signal components. Z2 and Z3 also function to keep undesired signals in the i-f passband from entering the i-f amplifiers.

CRYSTAL OSCILLATOR

A single oscillator located in the frequency multiplier-oscillator chassis (fig. 15-3, discussed later) serves as the basic frequency generating source for both the receiver (local oscillator) and transmitter of the AN/URN-3. The circuit of the oscillator (fig. 15-4) is that of a crystal-controlled oscillator with tickler coil feedback, and has a frequency output ranging from 40 to 50.5 mc. This frequency represents 1/24th of the final radio frequency output.

The crystal, Y1, is mounted on a constant temperature oven. Two indicator lights, I1 and I2, are mounted on the front panel of the frequency multiplier-oscillator chassis (fig. 15-3). When the equipment is initially turned on, the high and low indicator lights are energized. When the temperature of the oven has been raised to within a few degrees of its final operating temperature (75° C), the high heater and light are deenergized. The low heater will then maintain the oven at a constant temperature (75° C), as indicated by the low light blinking on and off as the low heater is alternately energized and deenergized. R4 and R5 are current limiting resistors for the heater circuit.

Meters M2 and M3 indicate the oscillator grid current. M2 is mounted on the front panel

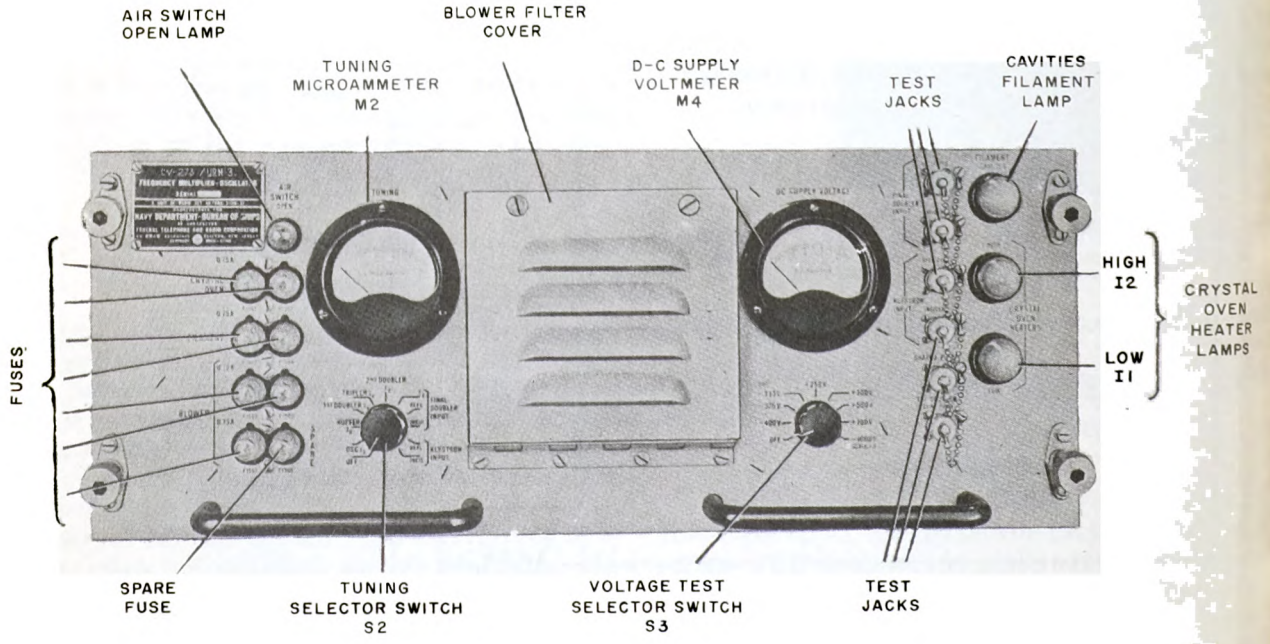

Figure 15-3.—Frequency multiplier oscillator, front panel view.

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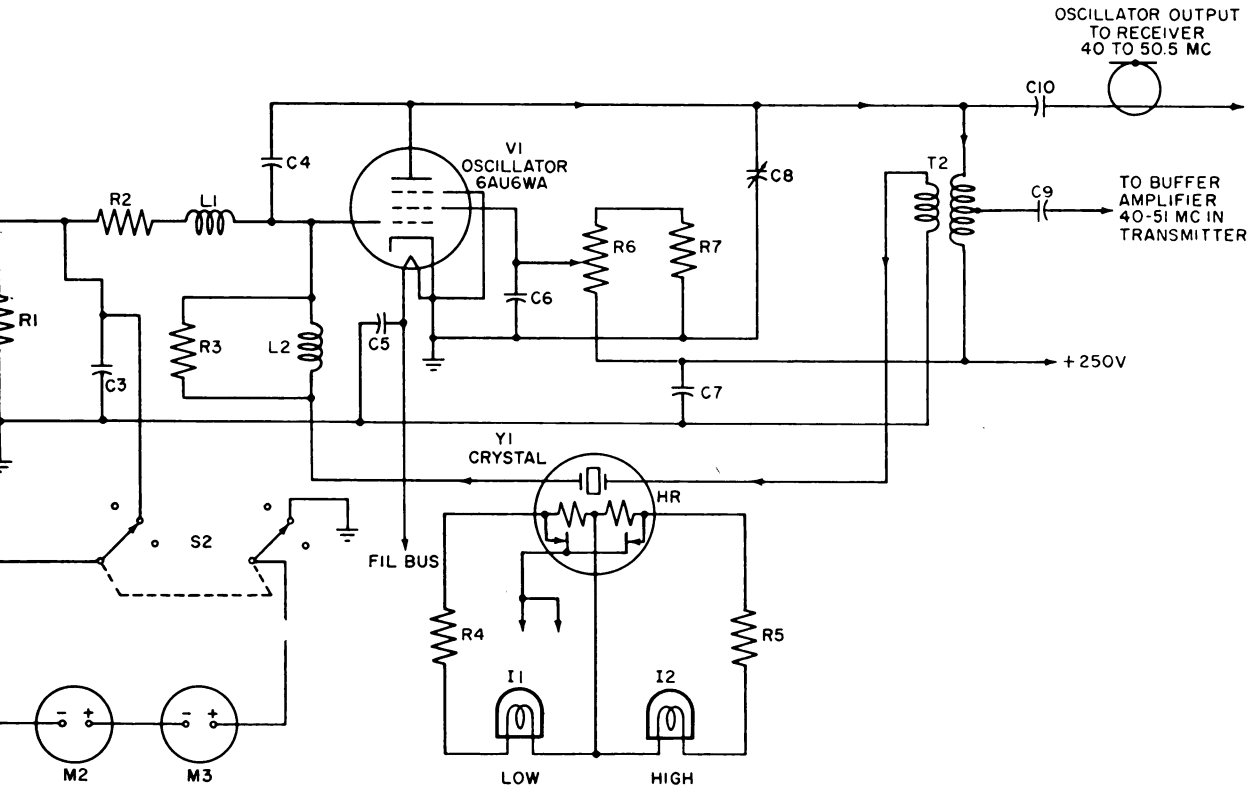

Figure 15-4.—Frequency multiplier oscillator, schematic diagram.

the frequency multiplier-oscillator chassis, hile M3 is mounted directly behind the panel r convenience during tuning operations. The eters indicate oscillator grid current when e meter selector switch, S2, is in the OS-ILLATOR position. R1 acts as a shunt to limit e current through the meter movement. C3 lters the oscillator grid signal to apply a more onstant potential across the meter.

R1 and R2, in conjunction with C3 and the rid-to-ground capacitance, form the grid leak las circuit for the oscillator. L1 is a low-esistance component to the grid bias current, ut presents a high grid-to-ground impedance o the oscillator signal. This action reduces e loading effect at the oscillator grid. C4 rovides a small amount of neutralizing voltage rom plate to grid (180° out of phase) to prevent purious frequencies not controlled by the os-illator crystal.

The plate tank circuit comprises C8 and the rimary of transformer T2 through which the B upply voltage to V1 is applied. The T2 sec-ndary acts as a tickler coil that provides posi-ive feedback through the crystal, Y1, and L2 and R3 (in parallel) to the control grid of the oscillator. The plate tank of a crystal controlled oscillator is normally tuned inductively. However, the phase shift across T2 necessitates the use of the phasing network (L2 and R3) to insure the proper phase of the feedback voltage. The amplitude of this signal is greater than that supplied by C4, and is therefore capable of sustaining the oscillations at the desired frequency.

Screen voltage is obtained from the arm of R6, which is a part of a voltage divider comprising R6 and R7, connected from the B supply to ground. Thus, the screen voltage can be adjusted to control the amplification factor of the oscillator tube, or to compensate for variations in transconductance with age. C5 bypasses to ground the oscillator and other high frequencies coupled through the cathode-to-filament capacitance. C7 bypasses these frequencies around the B supply.

The oscillator output is coupled by C10 to the amplifier and frequency multipliers in the receiver chassis (fig. 15-5), and from a tap on T2 to the buffer amplifier in the transmitter

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(discussed later). The signal amplitude applied through C9 to the buffer amplifier is reduced to prevent buffer grid current and loading of the input circuit of the transmitter.

FREQUENCY MULTIPLIERS (RECEIVERS)

The frequency multiplier section of the radio receiver multiplies the 40 to 50.5 mc, crystal-controlled oscillator frequency (from the frequency multiplier-oscillator chassis) to a frequency between 962 and 1024 mc, or between 1151 and 1213 mc. The lower frequencies correspond to low-band reception, while the higher frequencies are used for high-band reception.

The major reason for operating the receiver and transmitter from the same fundamental frequency source is to avoid the possibility of beat frequencies between the receiver local oscillator and the transmitter.

The oscillator frequency is applied to the receiver (fig. 15-5) via the primary of the input transformer, T3, The tuned secondary of T3 applies the signal through parasitic suppressor R10 to the grid of the first amplifier, V1. R8 and R9 serve as the grid return resistance to ground. A small current sometimes exists between the control and grid and cathode of a vacuum tube due to the dissimilarity of the metallic materials of the control grid and cathode of the tube. The dissimilarity between the elements is caused by the deposition of active material from the cathode or getter on the control grid. The small cathode-grid current (always in microamperes) is referred to as a contact current, and is usually conducted through a large grid-to-cathode resistance to develop a negative voltage at the grid. The potential produced between the grid and cathode (across the grid-cathode resistance) is called a contact bias potential.

A contact current is conducted through R8 and R9 (fig. 15-5) to provide the only bias voltage applied to V1. R9 also forms a part of the metering circuit, and will be considered later in a discussion of that circuit.

The output of V1 is a pulsating current at the same frequency as that of the input (40 to 50.5 mc). The output voltage is developed across L3. C14 couples the output to a shunt peaking coil, L4, which is variable to boost any frequency within its range as desired. This signal is coupled by C15 to the control grid of the tripler stage, V2.

Grid leak bias for V2 is established by C15, R12, and R13. The output tank comprising L6 and C19 is tuned to the third harmonic of the input (fundamental) frequency. The Q of the tank is such that three cycles are produced by the tank oscillations for each input pulse from the V2 plate, and frequency tripling is accomplished. The output tuned circuit of the tripler stage is adjustable over the range from 120.35 to 151.6 mc.

The tripler output is applied through C20 to the second amplifier control grid, V3. Bias is produced by C20, R15, and R16. Output tank L8-C25 is tuned to the frequency of the input.

L9 increases the grid-to-ground impedance at the V4 (doubler) input. Output tank L11-C31 is tuned to the second harmonic of its input to produce frequency doubling. The circuit is adjustable over the range from 240.5 to 303 mc.

The input to the quadrupler stage, V5, is taken from a tap on L11 to limit the signal amplitude applied at the grid. Cathode (safety) bias is adjustable by R26 in the cathode circuit to limit the amount of oscillator power fed to the mixer.

The quadrupler stage is connected as a grounded grid amplifier whose plate tank impedance is that of Z4. Z4 represents a short-circuited coaxial line, which is a quarter wavelength at the mean frequency of the band. C34 tunes the circuit to the fourth harmonic of the input frequency and quadrupling is accomplished. The Z4 output is adjustable by C34 over the range of the band.

A tacan station normally operates on a single frequency. All changes in frequency greater than 2 mc in final output frequency necessitate retuning the frequency multiplier. The tuning of the individual stages of the frequency multiplier is facilitated by the receiver panel meter and meter selector switch S3 (fig. 15-1). For example, when the switch is set at position 2, the meter is connected into the grid circuit of V1, and indicates maximum reading when the slug of T3 is correctly positioned. Likewise, when the meter selector switch is placed in position 3, the V2 (tripler) grid current is to be adjusted for maximum by positioning the slug of L4. The same principle is followed in tuning the remaining circuits of the frequency multiplier. R27 is permanently connected in series with the meter to limit the meter current.

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* ADJUSTABLE PRE-SET FOR 40 MICROSECONDS

** WHEN RECEIVER IS NOT RECEIVING DISTANCE INTERROGATIONS, WAVE SHAPE WILL BE ONLY NOISE UP TO STAGE V20

NOTES:
1. ALL RESISTORS ARE 1/2 WATT UNLESS OTHERWISE SPECIFIED.
2. CAPACITOR VALUES ARE IN μμf UNLESS OTHERWISE SPECIFIED.
3. INDUCTANCE VALUES ARE IN μh UNLESS OTHERWISE SPECIFIED.

Figure 15-5.—Radio receiver, schematic diagram.

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The output of the frequency multiplier section s coupled along a coaxial line to the crystal iodes in the balanced mixer circuits, as iscussed.

PREAMPLIFIER STAGES

Preamplifier stage V6 amplifies the 63-mc ignal coupled from the T1 secondary. The ube is a sharp cutoff pentode, which is connected as a triode. Plate inductor L23 is slug uned to facilitate alignment of the amplifier for relatively flat i-f response centered at 63.0 ic. The V6 output is coupled to the cathode of grounded grid amplifier, V7. The overall ircuit arrangement of V6 and V7 is that of cascode r-f amplifiers which provide about the ame gain as a single-stage pentode amplifier ut with considerably less noise.

The V6 output is applied through C48 to the athode of a grounded grid amplifier, V7. For ery high frequencies, the grounded grid amplifier removes the feedback coupling between grid nd plate, and places the grid plate interelectrode capacitance effectively in parallel with the oad. The input signal is introduced into the athode circuit in series with the grid-to-cathode oltage in the normal manner.

Bias for V7 is developed by R31 and C47. egenerative feedback is also established since he signal output is developed back at the V6 rid through the cathode conduction via the T1 econdary.

Coupling between V7 and V8 is accomplished y a TEE arrangement of series and parallel uned circuits. L24 and L26, with their associated distributed capacitance, form the series eaking circuit. These inductors contain adustable slugs, which vary the point of resonance eaking) of the circuit. L25 is a shunt peaking ircuit for the output. A similar coupling arangement is employed in the V8 output circuit, sing L27 as the series peaking coil, and L28 s the shunt coil. The V8 output is coupled long a 50-ohm line to the i-f amplifiers.

INTERMEDIATE FREQUENCY AMPLIFIERS

The receiver i-f amplifiers increase the trength of both the signal and noise from the reamplifiers to a level sufficient to operate Ferris discriminator. Echo suppression circuits are employed in the i-f amplifiers, which llow operation of the equipment in the vicinity f large reflecting objects.

The input circuit to the grid of V9 consists of the equivalent of a parallel-tuned, 50-ohm, line-to-grid coupling impedance formed by L30. Because of the impedance matching afforded by this circuit, the i-f response is substantially independent of the cable length between the preamplifiers and the i-f amplifiers.

The i-f voltage is amplified in tubes V9 through V13. Plate inductances L32 through L35 are stagger tuned to center the i-f response around 63 mc, and to provide the required i-f bandwidth of 2.7 ±0.3 mc.

Echo suppression is obtained by means of RC discharge networks in the grid circuits of V10, V12, and V13. In the presence of large signals, R44 and C68, R52 and C74, and R57 and C83, each develop sufficient bias voltage to reduce the gain of their respective amplifiers for a period proportional to the signal level input.

Squitter control in the form of a negative voltage from the video amplifier is fed to the grid of V9 and V10, thereby controlling the gain of these stages. The squitter control circuit is treated in detail later.

FERRIS DISCRIMINATOR

The Ferris discriminator (V14) detects the signal and provides adjacent and near-adjacent channel rejection. Low impedance points in the tank circuits (comprising C89, C90, and L37 and C93, C94, and L38, respectively) are used to feed the discriminator diodes. In this way, the tube capacitance is shunted across a smaller portion of the total tank impedance. Thus, the discriminator is relatively unaffected by variations in tube capacitance. In addition, the circuits are temperature compensated to maintain frequency stability over the temperature range from -54°C to +85°C (-65°F to +183°F).

The tank circuits of the discriminator are tuned to 63 mc. The circuit feeding V14A, however, has a Q approximately three times that of the tank circuit feeding V14B. The characteristic of the V14A tank circuit (for c-w reception) is shown at A of figure 15-6, while the characteristic of the V14B tank is shown at B of the figure.

V14A attempts to conduct a negative output across R66 and C97, while V14B attempts to develop a positive output across R65 and C96.

Since the Q of the V14A tank is higher than that of the V14B tank, the negative voltage developed through V14A conduction at the center

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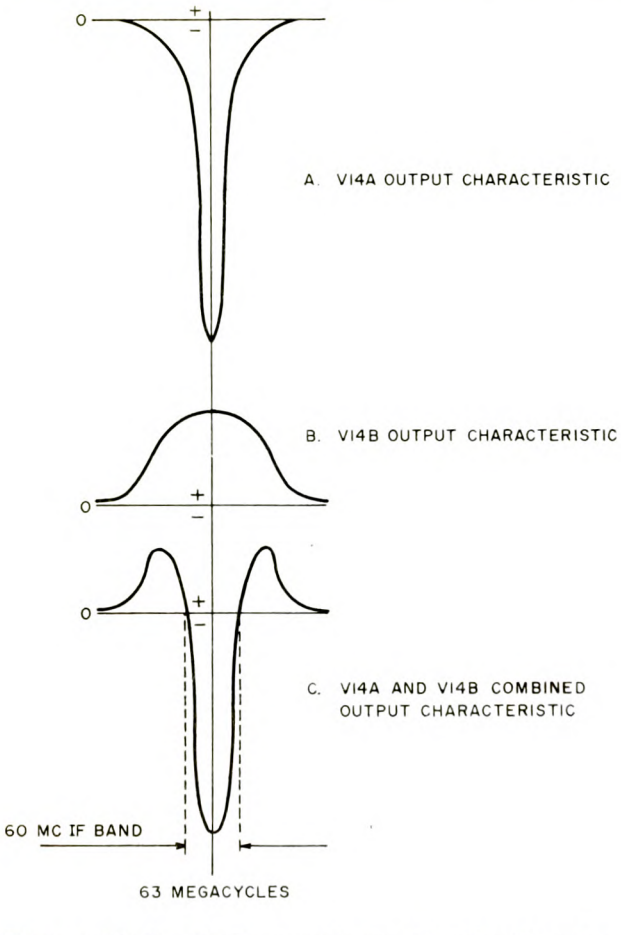

Figure 15-6.—Radio receiver, Ferris discriminator, c-w response curves.

of the i-f band is greater than the positive voltage developed through V14B conduction at the same frequency. The output to V15 is the algebraic sum of these two voltages. The normal output ot V15 within the 63-mc, i-f pass band is therefore a negative signal that is amplified normally in V15.

Note that a positive voltage appears on the V15A grid for adjacent channel frequencies. V15A is lightly biased as a result of low tube conduction through R69. When adjacent channel (positive) frequencies appear at its input, clamping diode, V16A conducts to quickly charge C98 negative at the grid side through C97 and C96. This negative voltage is sufficiently large to cut off V15A for the duration of the adjacent channel input.

The inverted output at the plate of V15A is applied through C100 to the grid of a similar circuit comprising V15B. Positive input signals allow V15B to conduct. Negative input signal are developed across a clamping diode, V16 to cut off V15B, thereby eliminating any re maining portion of the adjacent channel fre quencies.

The output of the Ferris discriminator an video amplifier consists of interrogation pulse pairs spaced 12 μs apart. The width of eac pulse is approximately 3 μs.

VIDEO CHASSIS

The video chassis decodes the interrogatio pulse-pairs and a sufficient number of nois pulses to maintain the receiver output at 270 ±90 pps. In addition, the video amplifier blanks itself out for 40 μs after each decoding of a pulse-pair, so that a 40-μs time minimum is maintained between pulses sent to the coder indicator.

The V15B output is coupled by C103 to the control grid of video amplifier V17. A portion of this signal is applied through C119 to the control grid of cathode follower, V22A, for purposes described later.

The V17 input is amplified and inverted, and coupled to blanking gate stage V18. The gating action of this stage and the amplitude and shape of the gate pulse will be considered later. It is sufficient at this time to state that V18 passes the video signal and the noise pulses.

The output of the blanking gate stage is coupled by C108 to the control grid of the final video amplifier, V19. Positive portions of the input are clipped by CR6. The plate output is applied via C109 directly to the suppressor grid of the decoder tube, V20, while the control grid signal is fed through a 12-μs delay line. It should be recalled that the interrogation pulse-pairs are spaced 12 μs apart.

V20 is a semiremote cutoff pentode biased to cutoff by a positive voltage obtained from the junction of R87 and R88, which form a voltage divider from the B supply to ground. Conduction of V20 requires the simultaneous application of positive voltage on both the control and suppressor grids.

When the first pulse of an interrogation pulse-pair, or the first positive pulse of a pair of noise pulses (separated by 12 μs) arrives at the suppressor grid of V20, it is not applied simultaneously to the control grid because of the12-μs delay line. Thus, V20 remains in the cutoff condition.

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After 12 μs have passed, the first pulse of ıe pair arrives at the control grid coinci-ɐntally with the arrival of the second pulse t the suppressor grid. V20 therefore conducts hen the second pulse of a pulse-pair is on the uppressor grid, and the first pulse is on the ontrol grid.

The characteristics of the decoder (V20) in-ure that only those pulses (noise or interroga-ion) having a spacing of 12 μs will be acted pon by the decoder circuit. It should be noted, owever, that since the decoder is biased to cut-ff, the pulse input must be of sufficient strength o cause conduction.

This means that not all decodable noise or nterrogation pulse-pairs will be decoded, but nly those above a minimum value. The re-eiver is designed to maintain the number of ecoded pulse-pairs at 2700 ±90 pulses per econd.

The decoder (V20) output consists of a ingle negative pulse for each decoded pulse-air. This pulse is used to trigger a one-shot nultivibrator, V21. V21A is zero biased and ormally conducts. V21B is held beyond cutoff y a positive voltage developed at its cathode y R96, which is a portion of a voltage divider omprising R96 and R98, connected from the 3 supply to ground. The negative input from 720 cuts off V21A to cause its plate voltage to ise, initiating the multivibrator action.

V21 provides three output signals. One out-ut (from the V21B plate) feeds the radio re-eiver blanking, one-shot multivibrator, V28. A second output is coupled to a pulse amplifier, 726, in the radio receiver squitter control cir-uit (discussed later). The third output feeds ne of two cathode follower output stages in tube 722. The output of V22B is a video pulse with n amplitude of about +30 volts and a duration f between 3 and 4 μs. This signal is the output of he radio receiver, which is applied to the video utput jack, J1, for connection by means of in-erunit cabling to the keyer circuits in the coder-ndicator. This output is also made available at est output jack, J2, for test purposes.

The second cathode follower, V22A, serves is the output stage to couple a portion of the ideo signal derived from the Ferris discrimi-ator as it appears before decoding at the input o the video chassis. The undecoded interroga-ion pulse-pairs and noise appearing at the utput of this cathode follower stage are made vailable at the "future use output" jack, J3, for application to special external equipment, when provided.

BLANKING CIRCUIT

The blanking circuit performs a dual purpose. It is responsible for the blocking of the decoder input circuit for 40 μs after a pulse-pair has entered the circuit, and for eliminating undesirable feedback from the klystron in the transmitter to the receiving circuits during transmission.

The B section of the blanking multivibrator, V28, is normally held beyond cutoff as a result of the positive voltage at the R121-R126 junction, which is applied to the V28B cathode. V28A normally conducts, since cathode bias alone controls the current through this tube section.

The negative-going output of the V21 B multivibrator plate cuts off V28A. The plate voltage rise of this tube section is coupled by C123 to the B section of V28, causing conduction. The positive output at the V28B cathode is coupled by C124 to the blanking phase inverter and isolation amplifier, V27. The pulse is amplified and inverted, and fed to the suppressor grid of blanking gate, V18.

It should be recalled that the second pulse of the input pulse-pair caused the triggering action at the V21 multivibrator. Likewise, the trigger to the blanking multivibrator, V28, and to the blanking gate stage, V18, corresponds in time with the second pulse of the input pair. The trigger pulse to V18 insures the cutoff of this stage for 40 μs from the leading edge of the second pulse of the pair.

Blanking time is adjustable by means of the "blanking time adjust" potentiometer, R127, which varies the time constant of the RC network to the multivibrator. The control has an adjustment range of between 20 and 45 μs, but is preset at the factory to produce a blanking time of 40 μs.

Note that V27 receives a high value of bias from the -105-volt regulated supply. Thus, conduction takes place only on strong positive pulse signals from the multivibrator.

SQUITTER CONTROL CIRCUIT

The output of the V21B multivibrator cathode, which is applied via C125 to the control grid of the pulse amplifier, V26, is used to derive the squitter control voltage. Squitter control involves the action of a circuit in its use of noise

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pulses to maintain the total pulse-pairs in the receiver at 2700 ±90 pps.

Grid limiting is permitted at the input to V26, which renders the output of this stage substantially independent of the amplitude of the input pulse. The output of V26 is transformer coupled by T4 to the pulse stretcher and counter diode, V25. As V25 conducts on negative input alternations, C128 charges negative toward the plate of the diode. C128 holds most of its charge during positive alternations at the V25 cathode, since the only discharge path is through the high resistance of R112. As a result of this action, the voltage on the grid of V24 is essentially a direct current. V24 is connected across both the negative and positive supplies. The cathode connects through R107 to the -105-volt source, while the plate connects directly to the +150-volt source.

The potential at the cathode of V24 is applied to the first and second i-f amplifiers as squitter control voltage. When the bias voltage on the grid of V24 is correct to effect a squitter rate of 2700 ±90 pps, the voltage at the cathode, because of the drop across R107, is approximately -5 volts with respect to ground. This -5 volts at the grid of the first and second i-f amplifiers controls the gain of these stages.

If the receiver output tends to fall below 2700 ±90 pps, the number of pulses fed into the pulse amplifier, V26, also decreases. This results in a lower d-c output from the pulse stretcher diode, V25, and a less negative bias on the grid of V24. Increased current flow in this stage results in a greater voltage drop across R107, and consequently a less negative voltage at the cathode with respect to ground. The reduced squitter control voltage to the i-f amplifiers results in a general increase in the level of the signal reaching the decoder. Thus, a greater number of decodable pulses attain sufficient amplitude to operate the decoder.

If the pulse output tends to rise above the requisite number (2700 pps), a greater number of pulses reaches the pulse amplifier, V26, and are coupled by T4 to the pulse stretcher and counter circuit. The more frequent conduction of V25 charges C123 to a higher value, and the bias at the V24 grid is increased. The decreased current through R107 increases the negative potential applied to the i-f amplifier grids, and the rise in the number of decodable pulses reaching the decoder is prevented.

Since without the application of the bias to the grid of V24 the voltage drop across R107 may result in a high positive voltage (in excess of +20 volts, d-c) applied to the i-f amplifiers, d-c clamp diode V23 operates to prevent the voltage from rising above +5 volts while the receiver is warming up. Without the clamping action of V23 the +20 volts applied to the grids of the amplifiers will completely block the receiver, and may cause possible damage to the receiver tubes.

As soon as the receiving tubes become heated to their operating temperature, pulses are decoded, and bias is applied to the V24 grid. The voltage drop across R107 (the squitter control voltage) becomes negative with respect to ground, and V23 ceases to conduct.

POWER SUPPLY

The radio receiver power supply provides regulated -105 volts d-c for the video amplifier, and regulated +150 volts d-c for the complete receiver. The B supply is electronically regulated by V30, V29, VR1, and VR2.

CODER-INDICATOR

The coder-indicator (fig. 15-7) receives its input from two sources within the radio beacon. These are the radio receiver and the antenna reference pulse generator. The radio receiver supplies the decoded interrogation pulses appearing at the video output connector, J1 (fig. 15-5). This signal, consisting of pulses with an amplitude of +30 volts and between 3 and 4 μs duration, is brought into the coder-indicator by way of input connector J3 (fig. 15-8), and is routed to the tube, V14. The coder-indicator processes these pulses and applies the necessary zero distance delay (described later) before feeding the pulses to the transmitter section of the radio beacon.

REFERENCE BURST CIRCUITS

The antenna reference pulse generator supplies the 15- and 135-cps trigger pulses (see chapter 14). These signals are brought into the coder-indicator by way of the input jacks, J2 and J1, respectively. The pulses are used to trigger the coder-indicator, which, in turn, generates 15- and 135-cps burst of pulses, which are essential elements of the bearing information transmitter to the aircraft.

The coder-indicator also generates an identification call signal of 1350 cps, phase-locked

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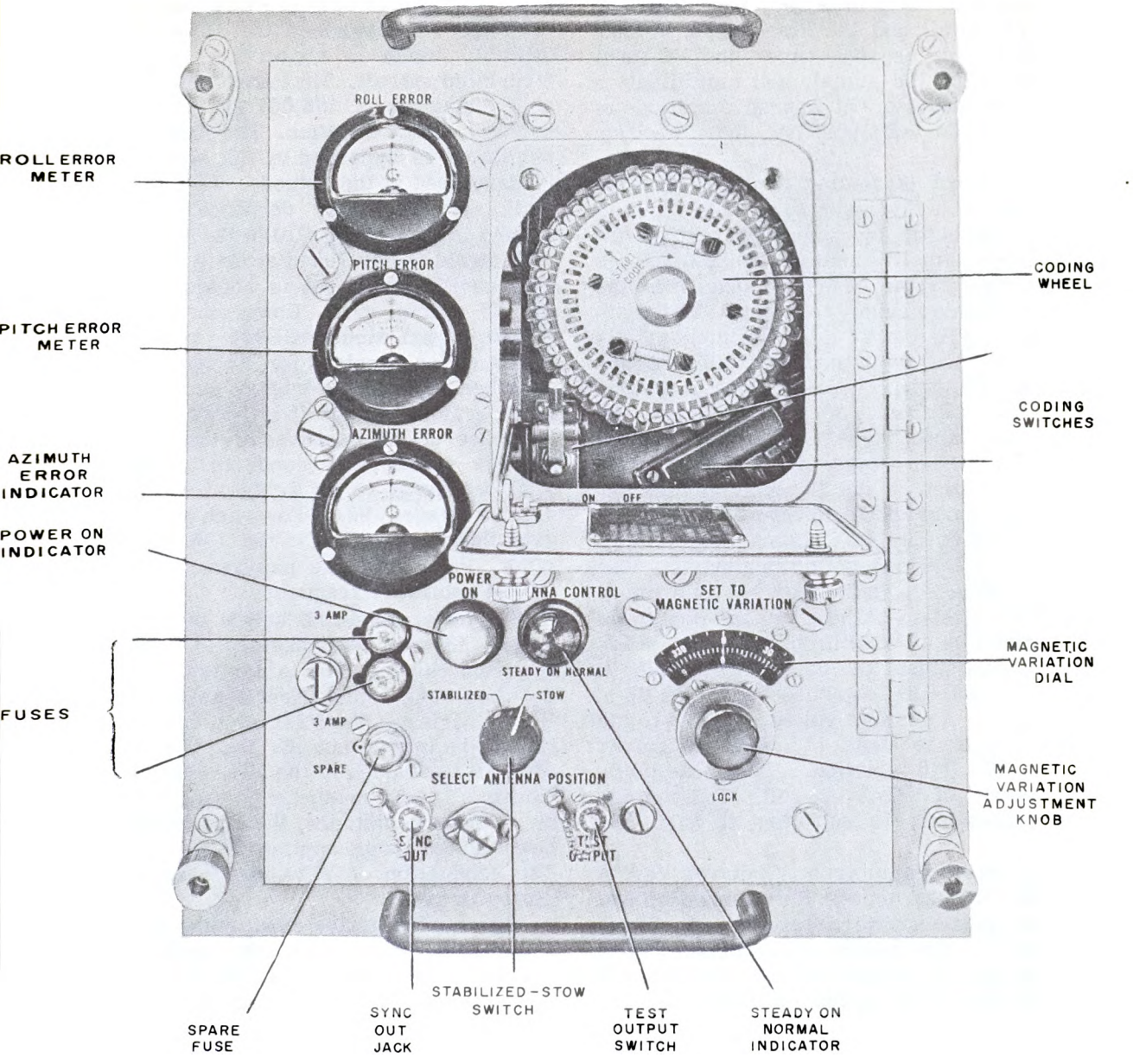

Figure 15-7.—Coder-indicator, front panel view.

with the antenna rotation, and applies international Morse code identification call keying to the transmitter at specific intervals of time. After assigning priorities to the components of the signal (namely, the reference burst, the identification call, and the distance interrogation pulses) the coder-indicator mixes the various components, and applies a 12-μs double pulsing to the entire output.

The 135-cps trigger pulse from the antenna is applied to a pulse amplifier, V1. The pulse has a peak-to-peak amplitude of about 10 volts, and is approximately 150 μs wide. Eight such pulses are applied to V1 for each 15-cps trigger pulse from the antenna applied to the V6 pulse amplifier grid.

The positive portion of the input pulse at the V1 grid causes grid current through R2 and

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R1. Thus, during this part of the input (grid limiting occurs) the grid is held at a relatively constant potential, and the plate output is constant. During the negative input, the V1 input is controlled by the signal, and amplification occurs. The output of V1 has an amplitude of approximately 60 volts positive, and a duration of about 30 μs.

The V1 output is applied through C2 to the control grid of a thyratron tube, V2, which is biased beyond cutoff by a negative voltage obtained from the R5-R7 junction. These resistors form a voltage divider, which is connected from the -105-regulated supply to ground. The positive trigger applied to V2 is of sufficient amplitude to cause the tube to conduct. The negative pulse produced at the V2 plate has an amplitude of about 100 volts and a duration of 75 μs. The width of the output pulse is increased by the amount of the time required for the V2 plate potential to return to the B supply voltage because of the charge of C3 through R9.

The V2 output (fig. 15-8) is coupled by C4 to the 135-cps pulse gate multivibrator, V3. V3A normally conducts a heavy current as a result of a positive grid-to-cathode voltage developed by the conduction of V3A through R11, the grid-to-cathode resistance of V3A, and R13 to the B supply. The cathode to ground voltage of V3 is about 30 volts. A positive voltage applied to the V3B grid, which is about +12 volts, is derived from the R15-R16 junction. Thus, the grid-cathode voltage of V3B is -18 volts. This voltage (during quiescence) is sufficient to hold V3B cut off.

The negative trigger from V2 drives V3A to cutoff. The voltage across R11 decreases and reduces the grid-cathode voltage of V3B. Thus, V3B conducts. The reduction in voltage across R11 comprises the output voltage, which is fed to the reference burst gating amplifier.

The length of the multivibrator output signal across R11 has its time constant (R13-C6) adjusted to be somewhat longer than the time required for the complete reference burst (144 μs). However, a second trigger pulse (described later) is applied to this circuit to terminate the burst at the appropriate time.

The positive-going gate voltage developed at the V3A plate is coupled through C5 to the suppressor grid of the 135-cps burst gate stage, V5. This tube is normally held cut off by the positive voltage developed at the R21-R24 junction. These resistors form a voltage divider, which is connected from the B supply to ground. This voltage serves as suppressor grid and control grid bias to hold V5 beyond cutoff.

Stage V4 comprises the 135-cps burst oscillator, which is a conventional crystal-controlled circuit. The burst oscillator operates at a frequency of 166.667 kc, with each cycle having a six μs duration. The plate circuit of the oscillator is bypassed by C8, so that the output is developed at the cathode. That portion of the total cathode output developed from R19 to ground is coupled by C10 to the V5 control grid. The signal is developed at the grid across R20. The oscillator output is constantly applied to the V5 control grid. However, the application of this signal alone will not produce the coincident action required for V5 conduction.

When the positive trigger pulse from the V3A plate arrives at the V5 suppressor grid (in the presence of the positive alternation of the oscillator signal), V5 conducts to pass the oscillator signal to an Eccles-Jordan frequency divider stage, V21. For each pulse from the gate multivibrator, V3, the 135-cps burst gate stage, V5, allows the passage of 24 cycles of the V4 oscillator frequency.

To understand the action of the frequency divider circuit, V21, assume that the B section is conducting and the A section is cutoff. The negative input signal (first negative pulse) from V5 appearing at the V21B grid causes this section of the tube to cut off. The plate voltage rise of V21B is coupled to the V21A grid. This action causes V21A to conduct a heavy current through the cathode resistance, thereby aiding the negative voltage being applied to the grid of V21B. The conduction of V21A increases, while V21B moves to cutoff.

The next negative input pulse applied at the V21A grid drives this tube section to cutoff, and V21B into conduction. Thus, it can be seen that the plate voltage swing of V21A has completed one negative pulse as a result of the two pulses from V5. The V21A output is applied to V22, which is the second of the two frequency divider circuits, each of which divides the frequency by two. The action of V22 is the same as that described for V21.

The frequency output of V22 is one-quarter the frequency of the crystal oscillator. Since the period of one cycle of the crystal oscillator output is six μs, the period for one complete cycle of the V22 output is 24 μs.

The reference burst circuit (V3-V4-V5, V21, V22, and V16) is designed to minimize time variations (jitter) between triggering

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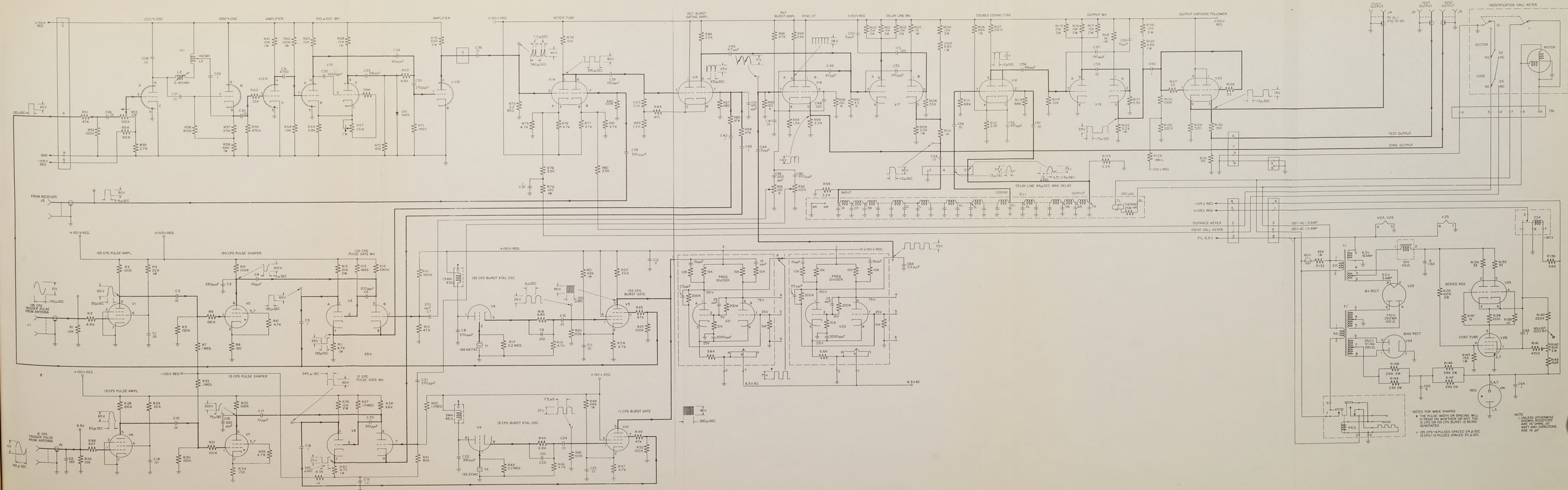

Figure 15-8.—Coder-indicator, schematic diagram.

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pulses from the antenna (J1 and J2 inputs) and the resultant reference burst (from V16). Since no synchronization exists between the triggering pulses from the antenna and the oscillator (V4) frequency, the triggering pulse and the first pulse of the reference burst (from V16) could vary slightly from burst to burst. To minimize this condition, the difference in time must be kept to within one-half of the repetition rate interval.

The reference burst oscillators V4 and V9 (fig. 15-8) are designed to operate at four times the frequency of the reference burst. In the case of the 15-cps burst, the oscillator operates at a frequency of 133.33 kc, having a period of 7.5 μs.

Assuming that the trigger pulse is received from the antenna at such time as to just miss the positive portion of one cycle of the oscillator, V4, the circuit must wait for the next cycle before the coincidence action of V5 can occur. Thus, the first useful cycle to enter the divider counting circuit, V21, is 7.2 μs late (at its maximum), which represents one-quarter the repetition rate interval of 30 μs.

In the case of the 135-cps burst, the oscillator operates at a frequency of 166.667 kc, and has a period of one cycle equal to 6 μs. This period represents the maximum jitter, which can occur between successive bursts of the 135-cps reference burst, and amounts to one-quarter the repetition period (interval between successive bursts) of 24 μs.

The frequency dividers in the reference burst circuits contain no provision for resetting. The counters (V21 and V22) will always be returned to normal with the proper number of pulses counted. For each 135-cps reference burst of six pulses, 24 oscillator pulses are allowed to enter the first divider circuit comprising V21. Since frequency dividers V21 and V22 divide by a total of four, any multiple of four pulses will return the circuits to their original condition.

In a similar manner, for each 15-cps burst, the frequency dividers, V21 and V22, divide by a total of four. For each 15-cps burst of 12 pulses, 48 oscillator pulses are allowed to enter the first frequency divider. Since this input is a multiple of four, the dividers will be returned to their normal condition at the end of the burst. The same divider is used for the division of both the 135- and 15-cps reference burst since the two bursts do not occur at the same time.

In the 135-cps reference burst circuit, after six pulses have been delivered by the frequency dividers, the action must be terminated by closing the gate being applied to V5 from V3A. The negative output at the reference burst amplifier, V16, (pin 6) is fed back through C45, R91, C7, and R15 to the control grid of the 135-cps pulse gate multivibrator, V3B. This tube section is normally nonconducting. However, when the gate is being applied to the 135-cps gate stage, V5, V3B is conducting. The negative pulses at the V3B grid are amplified and inverted by this tube section and coupled through C6 to the grid of V3A.

V3A is held cut off during the time that the gate pulse is being generated. The initial discharge of C6 through R13 is large, holding V3A far below cutoff, and the positive pulses from V3B are not of sufficient amplitude to cause the V3A conduction. As C6 continues to discharge, its discharging current decreases through R13, reducing the grid-to-cathode voltage. The rise in V3A grid voltage is therefore exponential, allowing the sixth pulse from V16 (via V3B) to cause V3A conduction, and the termination of the gate pulse to V5. R91 and C45 form a variable time constant network in the V16 coupling circuit to V3B, thereby controlling the time of arrival of the pulses at the V3B grid. In this manner, the time of the termination of the V5 gate pulse can be adjusted.

The 15-cps reference burst consists of 12 pulses, 30 μs apart. The 15-cps trigger pulse is introduced at J2. The pulse is amplified by V6 and, after the necessary peaking and pulse shaping, triggers the one-shot multivibrator, V8. The gating voltage from the multivibrator opens the V10 gate and allows the 133.33-kc oscillator, V9, to deliver 48 oscillator pulses to the frequency dividers, V21 and V22. These pulses are spaced 7.5 μs apart.

After division of the 48 pulses by four in the dividers, the proper spacing of 30 μs is available at the output of V22. Twelve pulses must be counted at the output of V22 before the action must be terminated.

PRIORITY CIRCUITS

The 135- and 15-cps gates coupled from the cathodes of the respective 135- and 15-cps pulse gate multivibrators are applied to priority tube V15 for the purpose of blanking the distance reply signals or the identification call signals during the reference burst periods.

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V15B is normally held beyond cutoff by the positive voltage at the cathode developed by the conduction of cathode follower, V15A, through the common cathode resistors, R83 and R84. Negative pulses from the keyer tube, V14 (discussed later), reduce the grid-cathode bias of V15A to cutoff, allowing V15B to conduct and pass the keyer pulses in the absence of the reference burst gate.

During the time of the reference burst, the decrease in the cathode voltage of V3 and V8, respectively, causes a negative voltage to be developed across R87 at the grid of V15B. These pulses are approximately -25 volts in amplitude. Negative signals from the keyer, V14, during this time cannot reduce the bias a sufficient amount to cause the V15B conduction. The output of V22 (reference burst frequency divider) is now being applied to the reference burst amplifier, V16, and will be the only input to the remaining circuits of the coder-indicator during the time that the reference burst occurs.

IDENTIFICATION CALL CIRCUITS

The coding wheel (see fig. 14-12, chapter 14) has secured around its edge metal segments, which may be extended after loosening the attaching screws. As the wheel rotates, the extended segments depress the roller of the keying switch, which is in the transmitter keying circuit. A motor (not shown) drives the wheel at 8 rpm, the speed required for properly keying (in International Morse Code) the radio beacon identification call. Each segment holds the keying switch closed for .125 second, which is the length of a dot. The length of a dash is .375 second, or the equivalent of three consecutive extended segments. There is a total of 60 segments on the coding wheel.

The coding wheel, driven by the motor at 8 rpm, completes a single revolution in 7.5 seconds. Since, however, the identification call is to be transmitted only every 37.5 seconds, the keying switch is cut out of the transmitter keying circuit for four revolutions of every five made by the wheel.

In the key-up position, the beacon receiver output, which consists of a total of 2700 distance interrogation and/or receiver noise pulses, is passed through the keyer tube, V14, to the reference burst amplifier, V15. In the key-down position, a 1350-cps identification signal is passed through the keyer tube, V14, to V15.

The 1350-cps oscillator, V11, receives the negative 132-μs gate from the 135-cps pulse gate multivibrator, V3A. This pulse occurs during the time that the reference burst pulses are being produced, and is used by V11 for phase-locking purposes only.

The V11 circuit is that of a one-shot multivibrator, using a series peaking circuit comprising L2, a portion of the inductor, L3, and C29. V11A is normally held cut off by negative voltage at its grid taken from the R54-R55 junction. R54 and R55 form a voltage divider connected from the -105-volt source to ground. V11B normally conducts because its only bias is that developed by the cathode resistor, R57.

When the positive trigger pulse from V3A arrives at the V11A grid, this section of the tube begins to conduct, with a corresponding drop in plate voltage. The discharge of C27 shock excites the circuit comprising L2, L3, and C29 into resonance at 1350 cps. This frequency is the 10th harmonic of the trigger frequency (135 cps). The slug in L2 is used to make the necessary frequency adjustments. R53 in the grid circuit is adjusted to lock the 1350-cps signal in phase with the gate pulse.

The oscillator output signals developed at the V11B cathode are coupled by C30 to the control grid of the amplifier, V12A. This input is shaped and amplified by V12A and applied to a 100-μs multivibrator comprising tube V13.

The circuit of V13 is that of a one-shot multivibrator. V13B is normally conducting since its grid and cathode are held at the same potential. The V13B conduction through R65 develops approximately 30 volts positive to ground. The grid of V13A is about 13 volts positive to ground as a result of a potential taken from the R62-R64 junction. (These resistors form a voltage divider from the +150-volt supply to ground.)

The grid of V13A is therefore held -17 volts to the cathode, and V13A is cut off.

When the oscillator output at the cathode of V11B is going less positive to ground, C30 will discharge through R59. This increases the negative voltage at the grid of V12A. The V12A output will be positive going to charge C31 through R64, thereby triggering V13A into conduction.

As the plate voltage of V13A decreases, C34 discharges through C35 and R71. This action applies a negative-going voltage (developed across R71) to the grid of V12B. During the same period, C32 discharges through R66 and R67 to increase the bias on V13B, resulting

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in the cutoff of V13B. The plate voltage rise of V13B (due to cutoff of this section) charges C33 through parallel paths of R70 and the low conducting resistance of CR1. Thus, at the time that the grid voltage of V12B is being driven negative by the V13A plate voltage decrease, CR1 effectively short circuits the grid of V12B for positive inputs.

At the end of the C32 discharge, V13B again conducts. The rise in cathode potential across R65 increases the grid-cathode potential of V13A and V13A again cuts off. The corresponding rise in plate voltage of V13A attempts to charge C34 through R69 and the low conducting resistance of CR1. However, the plate voltage of V13B is attempting to decrease since the tube is again conducting. The small positive voltage developed as a result of the CR1 conduction to charge C34 is immediately counteracted and overcome by the larger negative-going drop in the plate potential of V13B. C33 discharges through R69, C35, and R71 in parallel with R70. Thus, the voltage at the grid of V12B is again negative to the cathode.

From the above discussion, it should be seen that two negative pulses appear at the V12B grid for each complete cycle of the multivibrator. Therefore, the repetition rate of the pulses applied to the grid of V12B is twice the fundamental tone frequency of the oscillator, or 2700 cps. V12B amplifies and inverts these pulses and applies them to the keyer tube, V14.

The keyer tube determines whether distance reply signals or identification tone information will be transmitted during the time that the reference burst circuits are idle. Note that the input to the grid at pin 2 of V14 contains the identification call signals, and the input to the grid at pin 7 contains the distance reply signals from the radio beacon receiver. Only one of these signals may be transmitted at any one time. The function of the keyer tube is to select the signal to be transmitted at a specific time.

The cathode of V14A is held positive as a result of its connection to the R75-R76 junction. The cathode of the B section of V14 is also held positive in a similar manner through connection to the R77-R81 junction. Both of the voltage dividers at the cathodes are connected between the B supply and ground.

R78 and R79 connect the identification call line to the identification call keyer assembly, while R80 connects the distance information keyer line to the same unit. The connections are made from a terminal board, TB1. Terminal 4 of TB1 is grounded. Primary power is applied to the motor from terminals 5 and 7.

When the keyer motor is turned on, S1 is in the position shown. The position of S2 determines whether or not the signal at S3 will be effective in the circuit. The sector switch, S2, is operated by a cam (not shown), which rotates at one fifth the speed of the coding wheel. The letters, NC, at S2 indicate that the switch is in the normally closed position. Since the coding wheel operates S2 for four-fifths revolution to the NO contacts, a ground is placed on the lower end of R80, placing R80 and R81 in parallel at the V14B cathode. In the absence of this ground, one-half the B voltage is developed at the V14B cathode to hold this section of the tube beyond cutoff.

When the ground is applied to the lower end of R80, the V14B cathode voltage is reduced, and signals from the radio receiver can be amplified in this section and applied through C39 to the reference burst gating amplifier. These pulses are allowed to pass through V15 only when the negative gate pulse is not being received from either the 15-cps or 135-cps pulse gate multivibrator, as discussed.

For one-fifth rotation of the sector cam, S2 is in the NC position. During this period the code set on the code wheel operates S3 alternately to the NO and NC contacts.

During the interval when S3 is in the NO position, a ground is placed on the lower end of R79 to reduce the V14A cathode voltage, and this portion of the code is amplified. When the code wheel allows the contacts of S3 to return to the NC position, the ground is again placed on the lower end of R80 in the V14B cathode circuit and the distance reply pulses can be amplified. Thus, when S2 is in the NC position, both code (identification) and distance signals can be alternately amplified by V14A and V14B, respectively.

When the motor switch, S1, is turned off, terminal 3 of TB1 is grounded, and distance reply is continously transmitted. The code set on the coding wheel can therefore be changed without interrupting the transmission of distance reply signals.

As stated earlier, the B section of the reference burst gating amplifier, V15, is normally held beyond cutoff by the voltage developed across R83. Negative portions of the signals received from the keyer tube reduce the current through R83 (and therefore the bias) to the extent that V15B conducts. However, if the negative

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gate from either of the pulse gate multivibrators in the reference burst circuits is being received, V15B becomes more highly biased by the voltage developed across R87. In this condition, negative signals from the keyer tube are not of sufficient amplitude to cause the V15B conduction, and the reference burst pulses assume priority over the other signal elements. These pulses are received from the frequency divider, V22, and are applied to the control grid of V16A. The chosen output from the priority system is coupled through either C40 or C49 to the control grid of the delay line multivibrator, V17.

DELAY LINE AND DOUBLE CODING CIRCUITS

The reference burst gating amplifier (priority tube V15) feeds the identification call and distance reply signals to the delay line multivibrator, V17. The reference burst amplifier, V16, feeds the reference burst signals to V17. Since the various trigger pulses at the multivibrator are derived from separate sources, the multivibrator is needed to shape the mixed signal, and to make all of the applied pulses of equal length. The width of the pulse is important in the operation of a delay line.

V17B is normally held cut off by a positive voltage at its cathode taken from the R109-R105 junction. These resistors, with R104, form a voltage divider from the B supply to ground. The A section of V17 normally conducts since there is zero bias at its grid.

When a negative pulse is applied at the grid of V17A, this tube section is driven to cutoff. The plate voltage rise charges C51 through R108, making the V17 grid positive with respect to ground. The conduction of V17B through R109 causes a pulse of approximately 30 volts and 1.5 μs duration to be developed at the input to the delay line, DL1.

The delay line is constructed of phase-compensated L-C sections of one-quarter-μs delay each (except for sections involving special taps). The line consists of 175 sections. With all sections connected in series, the total delay to the input pulse is 44 μs from terminals 1R to 7R. It should be noted that the delay line is composed of four independent sections. The first section is included between terminals 1R and 2L, the second section between 1L and 4L, the third section between 3L and 6L, and the fourth between 5L and 7R. The total line is divided into these sections so that large sections of the line can be removed when not needed. This avoids undue reflections from the unterminated sections of the line.

The line is contained in a hermetically sealed can, which includes a thermostatically controlled heater. The drift in delay with temperature change is a maximum of 0.1 μs.

The delay line introduces the greatest portion of the zero distance, distance-reply delay in the radio beacon. The remaining delay (6 μs) is obtained from the other circuits of the equipment.

A secondary function of the delay line is to introduce a double pulsing action with a 12-μs spacing between the pulses. At the output of the delay line, two taps (6L and 7R, respectively) are provided 12 μs apart. The first pulse is applied to the grid of V18A from terminal 6L. The pulse from 7R is applied (12 μs later) to the grid of V18B. The plates of V18 are tied together. Both sections of the tube conduct, in turn, to couple two pulses through C56 for each pulse received at the delay line input.

The output multivibrator, V19, has its B section biased beyond cutoff by a positive voltage developed at the R122-R120 junction. Negative input pulses from V18 trigger V19A to cutoff. The rise in the V19A plate potential charges C58 through R121 to establish a positive voltage at the V19B grid. The circuit time constant allows V19B to conduct for 1.5 μs, developing a positive square-wave at the V19B cathode of approximately 55 volts. The circuit time constant is established by C57 and R114.

For each pulse received from the double coding tube, V18, the output multivibrator produces one output pulse. This pulse is coupled by C60 to the control grid of the output cathode follower, V20. The cathode follower is necessary to isolate the multivibrator and for impedance matching in the output circuit.

The two sections of V20 are connected in parallel to increase the power capabilities of the stage. The grids of V20 are biased beyond cutoff. The positive driving pulses from V19 cause conduction of V20, which produces the 1.5-μs positive pulses, approximately 15 volts peak, at the cathodes. These pulses appear at video output jack J4 for cable connection to the transmitter.

TEST CIRCUITS

A voltage divider comprising R130 and R131 in the cathode of V20 provides a test output of approximately one volt peak. This voltage is available at test output jack J5 on the front panel

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of the coder-indicator, and provides a point of examination for the coder-indicator output.

A portion of the frequency divider output fed to V16A is coupled from the cathode through C48 to the control grid of V16B. This section of the tube is connected as a cathode follower which develops the positive reference burst pulses from V22 at its cathode. The voltage across R94 is made available at the sync output jack, J6. The signal corresponds to the reference pulses present at the input to V17, and is a source of synchronizing voltage for an oscilloscope used to examine and count the reference burst. The V16B cathode follower arrangement is used to isolate the synchronizing voltage output from the main signal path.

POWER SUPPLY

The power supply (fig. 15-8) for the coder-indicator employs a single power transformer, T1, to supply both the plate and bias voltages. V23 constitutes a full-wave rectifier for the plate voltage, while V24 supplies the bias voltage. The B supply is electronically regulated by VR1, V26 and V25.

AMPLIFIER-MODULATOR

LOW-LEVEL MODULATOR

A front panel view of the amplifier-modulator showing the klystron oscillator is illustrated in figure 15-9. The input to the amplifier-modulator (fig. 15-10) consists of 3600 pulse-pairs, spaced 12 μs apart, arriving from V20 in the coder-indicator (fig. 15-8). These pulses are all of approximately 1.5 μs and have an amplitude of about 15 volts peak.

As a matter of simplification, it will be assumed that a single pulse is used as the modulator input. If this pulse is applied at a repetition rate of 7200 pps, the average pulse duty cycle will be the same as with the actual "full-traffic" input.

The video input pulses from the coder-indicator are received at J1 (fig. 15-10). The initial operation performed on the input pulse is a time delay accomplished in delay line DL1. The delay line is composed of lumped constants, comprising inductance, capacitance, and resistance. The artificial line has characteristics similar to that of an actual line; that is, it has the ability to delay the output signal with respect to the input signal. The longer the line, the greater will be the delay.

DL1 consists of 14 sections of .15-μs delay each, for a total delay of 2.1 μs. By tapping the line at appropriate points, specific delays can be obtained. Thus, the line can be adjusted to introduce different portions of the total beacon delay of 50 μs. The delayed pulse from DL1 is coupled by C1 to the control grid of the clipper-amplifier stage, V1A.

V1A represents an amplifier with an adjustable bias. The bias is obtained from a resistive divider consisting of resistors R19, potentiometer R20, and R21, which are connected in series from the -375-volt supply to ground. The bias voltage obtained from the arm of R20 is applied to the grid of V1A through grid leak resistor R1. This bias can be adjusted by R20 to obtain a specified amplitude output pulse at the plate of V1A for a wide range of input pulse amplitudes at the grid.

The negative-going output pulse of V1A is coupled by C4 to the control grid of V1B. This tube section operates at zero bias. The input pulse from V1A is of sufficient amplitude to drive V1B beyond cutoff, thereby limiting the amplitude of the output pulse.

The positive pulse at the plate of V1B is applied to delay line DL2, which acts as the plate load for this stage. A portion of the plate output is coupled by C5 to the grid of a cathode follower, V2A, for purposes that are discussed later.

DL2 has an input impedance of 1500 ohms, and a maximum time delay of 10.05 μs. The line is composed of 70 cascaded, low-pass filter sections. The delay line is terminated at its output end by a half section paralleled by an external 1500-ohm resistor, R14. The input end is terminated by a half section of the line in parallel with the output impedance of V1B and the input impedance of V2A.

The pulse widener, V3, consists of a series arrangement of V3A and V3B. The grid of V3A is held positive by a voltage taken from the arm of R15. This potentiometer is connected from the B supply to ground.

When the equipment is initially set into operation, V3A conducts momentarily through CR1 to charge C10. Both the voltage across C10 and the V3A cathode voltage rise exponentially until V3A is cut off.

V3B is held beyond cutoff by a negative voltage taken from the arm of R23. This potentiometer forms a portion of a voltage divider, comprising R22 and R23, which is connected from the -375-volt supply to ground.

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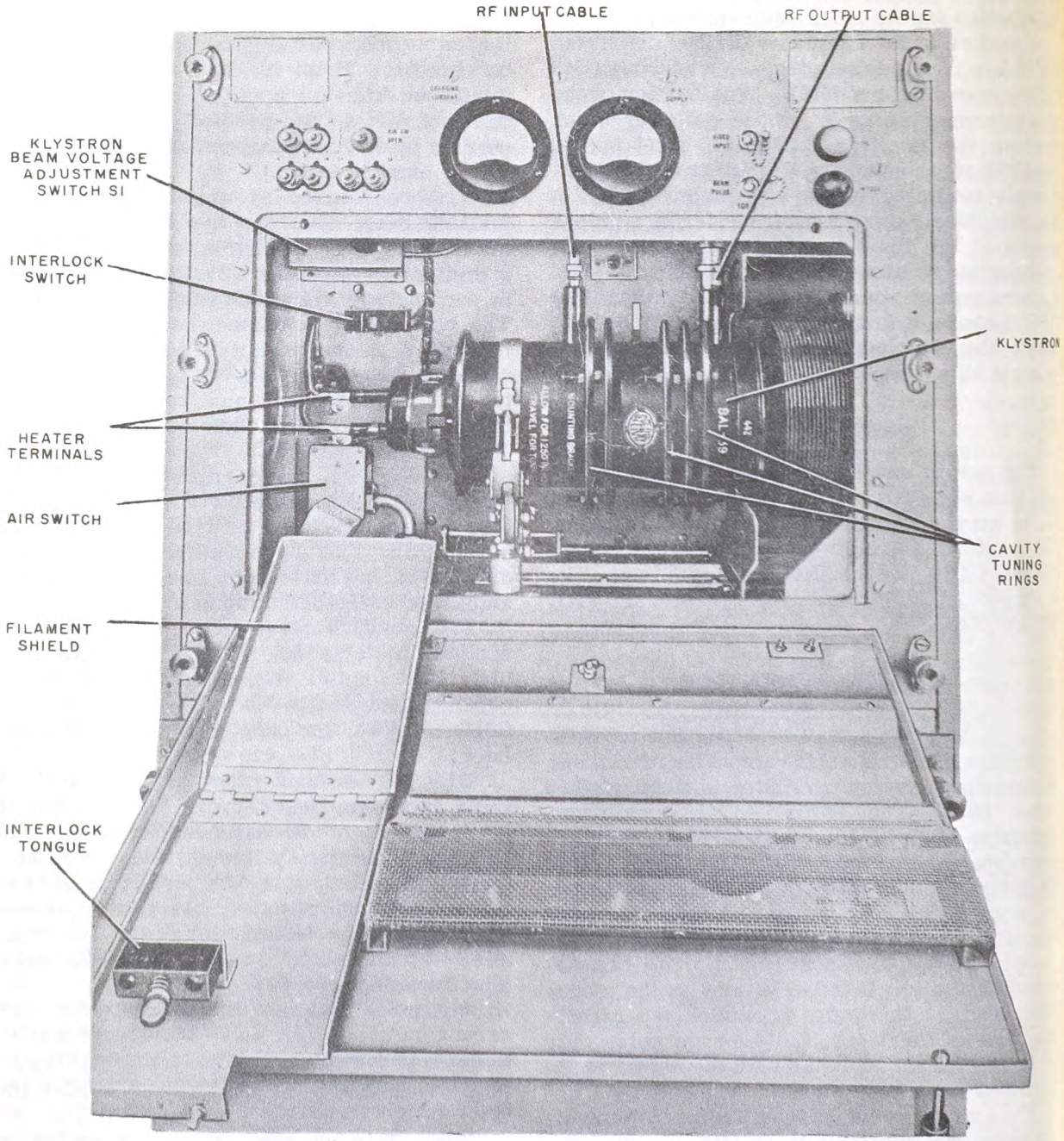

Figure 15-9.—Amplifier-modulator with klystron access door open.

When the first positive pulse arrives from the delay line to the grid of V3A, this section of the tube again conducts to further charge C10 to a more positive value. The resistance in the charge path includes the conducting resistance of CR1, and the conducting resistance of V3A. Both of these resistances are small so that C10 quickly charges with an increase in voltage of approximately the value of the pulse appearing at the grid.

At the end of the pulse input at the V3A grid, C10 has a very slow discharge path through C13,

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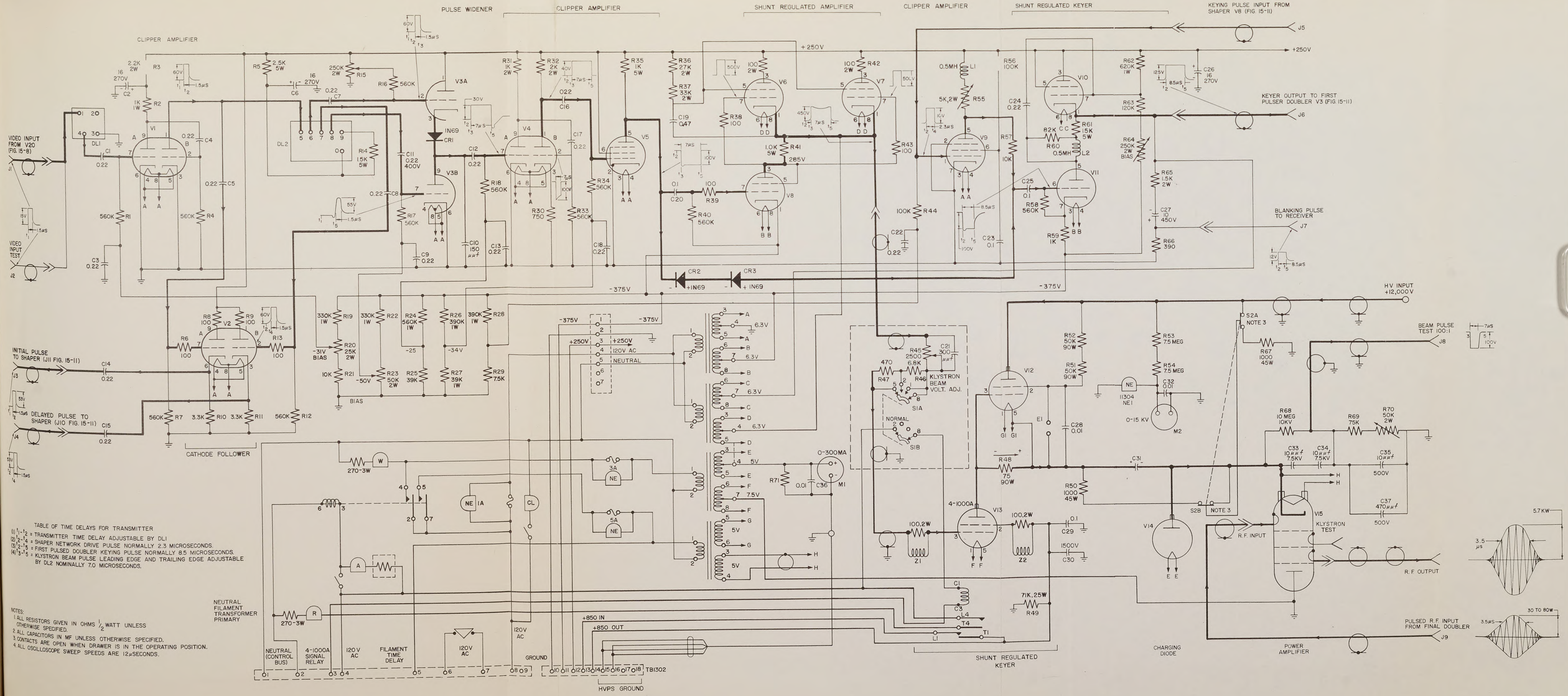
PULSE WIDENER
CLIPPER AMPLIFIER
SHUNT REGULATED AMPLIFIER
CLIPPER AMPLIFIER
SHUNT REGULATED KEYER
KEYING PULSE INPUT FROM SHAPER V8 (FIG. 15-11)
KEYER OUTPUT TO FIRST PULSER DOUBLER V3 (FIG. 15-11)
BLANKING PULSE TO RECEIVER
VIDEO INPUT FROM V20 (FIG. 15-8)
VIDEO INPUT TEST
INITIAL PULSE TO SHAPER (JII FIG. 15-11)
DELAYED PULSE TO SHAPER (JIO FIG. 15-11)
CATHODE FOLLOWER
+250V
-375V
-31V BIAS
-50V BIAS
-25
-34V
120V AC
NEUTRAL
KLYSTRON BEAM VOLT. ADJ.
NORMAL
0-300MA
HV INPUT +12,000V
BEAM PULSE TEST 100:1
R.F. INPUT
R.F. OUTPUT
KLYSTRON TEST
CHARGING DIODE
POWER AMPLIFIER
PULSED R.F. INPUT FROM FINAL DOUBLER
5.7 KW
30 TO 80W
NEUTRAL FILAMENT TRANSFORMER PRIMARY
NEUTRAL (CONTROL BUS)
4-1000A SIGNAL RELAY
120V AC
FILAMENT TIME DELAY
120V AC
GROUND
+850 IN
+850 OUT
HVPS GROUND
SHUNT REGULATED KEYER
TABLE OF TIME DELAYS FOR TRANSMITTER
(1) 1-2 = TRANSMITTER TIME DELAY ADJUSTABLE BY DLI
(2) 2-4 = SHAPER NETWORK DRIVE PULSE NORMALLY 2.3 MICROSECONDS.
(3) 2-5 = FIRST PULSED DOUBLER KEYING PULSE NORMALLY 8.5 MICROSECONDS.
(4) 3-5 = KLYSTRON BEAM PULSE LEADING EDGE AND TRAILING EDGE ADJUSTABLE BY DL2 NOMINALLY 7.0 MICROSECONDS.
NOTES:
1. ALL RESISTORS GIVEN IN OHMS 1/2 WATT UNLESS OTHERWISE SPECIFIED.
2. ALL CAPACITORS IN MF UNLESS OTHERWISE SPECIFIED.
3. CONTACTS ARE OPEN WHEN DRAWER IS IN THE OPERATING POSITION.
4. ALL OSCILLOSCOPE SWEEP SPEEDS ARE 12μSECONDS.

Figure 15-10.—Amplifier-modulator, schematic diagram.

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R18, and C12. Thus, only a small portion of the pulse amplitude is lost before the delayed pulse from DL2 overcomes the bias at V3B, to allow its conduction and the complete discharge of C10 through V3B.

The width of the pulse developed at C10 is determined by the distance (in time) between the leading edges of the two pulses from the delay line to V3A and V3B. This time is adjusted (at the delay line DL2) to 7.0 μs.

CR1 (fig. 15-10) prevents the negative portion of the output pulse at C10 from being coupled to the V3A cathode. The negative portion of the pulse at the output of C10 occurs because at the time of the V3B conduction, C10 is allowed to discharge to the value of the voltage from the plate of V3B to ground. When V3B is again cut off by the bias at its grid, C10 again charges to the more positive cutoff voltage existing from the V3A cathode to ground. The positive output of the pulse widener circuit is coupled by C12 to the grid of the clipper-amplifier stage, V4.

The A section of the clipper-amplifier tube, V4 (fig. 15-10) is biased at cutoff by a negative voltage obtained from a voltage divider comprising R24 and R25, which are connected from the -375-volt supply to ground. Thus, the less positive portions of the input pulse are not allowed to pass through the tube.

The clipped negative output at the V4A plate is coupled by C17 to the control grid of V4B. This section of V4 normally conducts because it is cathode biased by R30. The input pulse is of sufficient amplitude to drive V4B beyond cutoff, and clipping of the signal is accomplished. Degeneration is permitted at the cathode by R30 to improve the square shape of the output pulse.

The output of V4B is coupled by C16 to the control grid of amplifier V5. This tube is biased at cutoff by a negative voltage taken from the R26-R27 junction. The positive signal from V4 causes V5 to conduct, producing a negative pulse at its plate, which is coupled by C20 to the grid of V8.

V8 is connected in series with the parallel-connected tubes, V6 and V7, between the -375- and +250-volt supplies in a shunt-regulated amplifier circuit arrangement. This circuit develops a large output voltage with good output regulation, despite variations in load impedance.

The plates of V6 and V7 are connected to the +250-volt supply. The screens of the tubes are tied together and connected through R36 and R37 to the +250-volt supply. The voltage at the screens with respect to the cathodes is equal to the potential across C19. The cathodes of V6 and V7 are connected together and coupled to the plate of V8 through R41. The cathode of V8 is directly connected to the -375-volt supply. V8 is connected as a triode.

In the absence of a pulse from V5, a quiescent current flows through the series parallel combination of tubes of such value that the total voltage drop across the parallel connection of V6 and V7, in series with R41 and V8, is equal to the sum of the magnitudes of the positive and negative supply potentials. The total difference in potential from the V8 cathode to the B supply is 625 volts.

The grid of V8 is connected through parasitic suppressor R39 and grid resistor R40 to the cathode. Thus, zero bias is applied to this tube. The bias voltage of V6 and V7 is equal to the quiescent voltage drop across R41. The potential at the cathodes of V6 and V7 is approximately -240 volts. This potential is applied (during quiescence) through resistors to the grid of V13 (discussed later). The voltage is large enough to keep the grid of V13 beyond cutoff for its existing plate and screen voltage value (approximately 12,000 volts) in the absence of a driving pulse at its grid.

The negative output pulse from V5 at the control grid of V8 drives the tube to cutoff. The V8 plate voltage rises. The voltage drop across R41 drops to zero, and the potential at the grid of V13 rises. Since the voltage drop across R41 is zero, the bias applied to V6 and V7 is also zero. The path of conduction through V8 is now open, isolating the negative supply. The cathode conduction path for V6 and V7 is then through R45, the 5-8 contacts of S1A (in the position shown), through Z1, through the grid to cathode resistance of V13, and through the center tap of the "F.F" winding on the power transformer to ground.

HIGH-LEVEL MODULATOR

The circuit of the high-level modulator consists of a series arrangement of V12 and V13. The plate of V12 is connected to the high voltage supply (+12,000 volts). R51 and R52 are the screen resistors for V12, which are bypassed by C28. The screen voltage for V13 is obtained from a +850-volt supply through relay K1. Z1 and Z2, connected in the grid and screen circuits of V13, are parasitic suppressors.

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The klystron tube, V15, operates as a three-cavity amplifier, depending on the changes introduced in the velocity of a stream of electrons by alternately slowing some of the electrons and accelerating others, and using the transit time between the two occurences to produce the alternating current output. The electron beam of V15 follows the space-charge focusing principle, as discussed in chapter 14 of this training course. The beam follows a trajectory determined by the initial convergent angle at the electron gun and by the mutual space-charge repulsion forces of the electrons. The beam comes to a minimum and then spreads out, so that it does not have a uniform diameter throughout its length (see fig. 14-13, chapter 14). The pulse lengths used at the klystron are too short to allow enough ionization to occur in the tube to neutralize the beam space charge. The klystron has three cavities, which are tuned to the proper frequency by changing the gap spacing. The over-all power gain of V15 is well over 100 at its operating range of approximately 1,000 mc.

The cathode of the klystron is held at the potential developed at the C31-V14 junction. Before a pulse arrives from the shunt-regulated amplifier, V13 is nonconducting, and the voltage developed across R48 is zero. The bias voltage applied to V12 will also be zero, and C31 charges to approximately 11,000 volts through V12, and diode V14. The screen capacitor, C28, charges to the value of the screen voltage. Both capacitors (C28 and C31) are shorted by the contacts of S2 when the drawer is opened to protect the operator.

The a-c load on the high-level modulator, V13, consists of the klystron, V15, which is coupled through C31. The diode, V14, presents a very high impedance during the beam conduction of the klystron, so that the load on the modulator is the cathode-to-ground impedance of the klystron.

When a positive pulse is applied to the grid of V13 from the low-level modulator, the pulse drives the grid into the region of grid current. Thus, the impedance of V13 changes very rapidly from a very high value to a very low one. This impedance change takes place within the rise time of the input pulse.

During the period of the pulse, the voltage from the plate of V13 to ground must decrease. Thus, C31 discharges through the cathode to ground impedance of the klystron, V15. On the discharge cycle, the capacitor delivers approximately 10 percent of its fully charged voltage through the klystron. This represents the klystron beam pulse.

During the period of the pulse the voltage developed across R48 by the peak current discharged from C31 is sufficiently negative toward the grid of V12 to hold this tube beyond cutoff. Since there is no current flowing through V12 during the time of the pulse, the plate voltage of V12 is approximately 12,000 volts.

The discharge of C31 during the time of the pulse has the same effect as applying a negative pulse at the cathode of the klystron. This action causes the flow of beam current in the klystron. The time constant of the capacitor discharge path is such that for the longest group of pulses (the 24 pulse reference burst group) the voltage lost across C31, due to the loss of charge during the klystron conduction, is less than 10 percent of its initial charged value.

During the time of the pulse, the cathode voltage of V12 decreases with the plate voltage of V13. Since the voltage across C28 cannot change instantaneously, the screen-to-cathode voltage of V12 remains approximately the same. The screen-to-ground voltage, however, decreases. When this happens, C28 begins to charge toward the high voltage potential through V13, R48, R51, R52, and the resistance of the high-voltage supply, thereby increasing the screen-to-cathode voltage.

As the pulse decays to zero at the grid of V13, the voltage at the plate of the tube rises, and the impedance of V13 becomes very high. As the V13 current stops flowing the voltage across R48 drops to zero, and the grid-to-cathode voltage of V12 decreases to zero. V12 begins to conduct. The screen-to-cathode voltage of V12 does not change initially, but, as a result of the plate-to-cathode voltage decreases, the screen current will increase. The initial increase of screen current is supplied by C28.

When the driving pulse decays, C31 is connected to the high-voltage supply through the conducting path of V12 and the charging diode, V14. C31 regains its charge of approximately 11,000 volts. The low impedance of the diode minimizes the tendency for positive overshoot at the trailing edge of the pulse, and provides a low-resistance charging path for C31.

A meter, M1, measures the charging current through V14. C36 is a filter capacitor for the meter, while R71 is a thyrite resistor used to

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protect the operator or technician in case the meter becomes an open circuit.

A spark-gap, E1, is placed across C28 to prevent excessive voltages from destroying the capacitor. The high voltage is monitored by meter M2 in series with multiplier resistors R53 and R54. The meter is calibrated directly in kilovolts. The meter is shunted by a neon bulb, NE1, which becomes conductive in case of an open circuit through the meter, and thus protects the operator by preventing excessive voltage.

In order to monitor the beam pulse applied to the klystron, a frequency compensated voltage divider comprising R68, R69, and R70 is used. The ratio of the voltage division across R68 relative to that appearing across the series combination of R69 and R70 is approximately 100:1 This ratio of voltage division is independent of frequency since the time constant made up of C33, C34, and R68, is equal to that of C37, C35, R69, and R70. The voltage ratio is also approximately equal to the inverse of the corresponding capacitance ratio. The divider is tapped at the junction of R68 and R69, allowing this portion of the beam pulse to be developed at the beam pulse test jack, J8. This jack is located on the front panel of the amplifier-modulator chassis.

Switch S1 is the klystron beam voltage attentuator switch and is used when installing new klystron tubes. Because of high gaseous conditions usually caused by inherent properties of the tube and the fact that complete evacuation of the tube is not possible during manufacture, it is necessary to operate a new or idle klystron for approximately 20 minutes to bring the tube characteristics to normal. To prevent ionization of gas, which could cause great damage to the tube, a procedure referred to as "baking" is undertaken to drive the gas molecules back into the wall of the tube before full beam voltage is applied. The procedure is to drive the klystron with reduced beam voltage as follows.

When S1A is placed in a position so that its contacts, 1 and 8, are making, R46 and R47 are connected in series with the pulse signal to V13. This limits the drive at the grid to nearly zero volts. In this position the klystron beam voltage is 1/3 normal.

When contacts 2 and 8 are making, R46 is shorted, and a greater portion of the input pulse is applied to the V13 grid. In this position the klystron beam voltage is 2/3 normal.

When S1B is placed in any position except NORMAL (position shown in figure 15-10), the circuit to the coil of relay K1 is opened. The screen voltage of V13 is applied through the closed contacts of this relay, and is removed when the contacts are open. The screen of V13 is grounded through R49. In this condition (screen voltage removed), the plate to ground voltage of V13 is increased. During the beam pulse, the conduction path for the klystron is through the increased resistance of V13, through R48, and through C31 to the klystron cathode. The voltage across the klystron is therefore reduced during the "baking" process.

PULSE KEYING CIRCUIT

Tubes V9, V10, and V11, (fig. 15-10) comprise the pulse keying circuit for the first doubler stage in the frequency multiplier oscillator chassis. A positive pulse from V8 (fig. 15-11) in the shaper chassis of the frequency multiplier-oscillator is applied to the control grid of V9 (fig. 15-10). The grid is biased beyond cutoff by a negative voltage obtained from the junction of R28 and R29. These resistors are connected between the -375-volt supply and ground.

The leading edge, t_2, of the positive input pulse to V9 corresponds in time to the zero time delay pulse from delay line, DL2, which is applied to the shaper pulse-widener, V7A, (fig. 15-11) through cathode follower V2 (fig. 15-10). The trailing edge, t_4, of the pulse to V9 corresponds in time to the delayed pulse from tap 7 on DL2, which is also applied to the shaper pulse-widener, V7B, via cathode follower V2B. The a-c plate load of V9 is essentially the variable resistor, R55, in series with a peaking coil, L1. The purpose of L1 is to improve the high frequency response at the V9 output, thereby preserving the rise time of the input pulse.

The positive input pulse at the V9 grid causes the tube to conduct. The negative 2.3-μs output pulse is coupled through isolating resistor, R57, and through C25 to the grid of V11. R35, in the plate circuit of V5, is not affected by the a-c component of the V9 output because the high back resistance of CR2 and CR3 (for negative outputs of V9) isolates the two circuits.

The negative-going output pulse at the plate of V5 is coupled through the conducting diodes, CR2 and CR3, and through C25 to the grid of V11. Thus, two pulses appear at this grid. The

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arrival of the 2.3-μs pulse from V8 (fig. 15-11) in the frequency multiplier-oscillator corresponds, in time, to zero relative to the outputs of the delay line, DL2 (fig. 15-10). The leading edge of the 7-μs klystron beam pulse at the grid of V11 is delayed 1.5 μs relative to zero time at DL2. Therefore, the negative pulse at the V11 grid keeps the tube beyond cutoff for a total period of 8.5 μs.

During the time between pulse groups, the grid-to-cathode voltage of V11 is zero by a direct connection of the grid resistor, R58, to the cathode. The cathode is connected (through R59) to the -375-volt supply.

The d-c voltage at the cathode of V10 is directly applied to the control grid of the first doubler, V3 (fig. 15-11). This potential is used to supply keying voltage and bias for the doubler stage.

The d-c grid potential of V10 is supplied from a voltage divider comprising R62, R63, and R64, which is connected between the full values of the negative and positive voltage supplies. The grid voltage is obtained from the junction of R62 and R63, and can be adjusted by R64. Each change in the grid voltage produces a proportional change at the V10 cathode (with respect to ground) and likewise at the grid of V3 in the first doubler (fig. 15-11).

The cathode potential of V10 is determined by the quiescent current, which flows through the internal resistance of V11, peaking coil, L2 (paralleled by R60), and R61. The V10 cathode voltage developed as a result of this current is approximately -127 volts. C27, in series with R65 and R66, is charged with the upper plate negative with respect to ground.

The negative input pulse to V11 drives this tube to cutoff. Since the current through V11, L2, and R61 decreases to zero, the grid-to-cathode voltage of V10 also decreases. The voltage induced in coil L2, due to the decrease of current flowing in the circuit, is of such polarity and magnitude that during the rise time of the V11 input pulse, the voltage applied to the grid of V10 will be large enough to make the grid positive with respect to the cathode. The internal resistance of V10 then becomes very small, and the tube conducts. The magnitude of the voltage developed by the coil, L2, determines the rate of rise in the V10 conduction, and therefore the rise characteristics of the output pulse.

During the time of the V11 cutoff, the cathode conduction path for V10 is through R65, C27, and R66 to ground. One time constant of this circuit is approximately 20 μs. The total width of the driving pulse is 8.5 μs. Thus, the duration of the conduction period of V10 is short compared with the charge time of C27.

The voltage developed from the top of R66 to ground is approximately 125 volts and has a duration of 8.5 μs. This voltage appears at J6 and is coupled to the first doubler stage in the frequency multiplier-oscillator.

A positive 12-volt, 8.5 μs pulse used to blank out the receiver during transmission is taken from the junction of C27 and R66. This signal is connected through a cable to J4 (fig. 15-5) and is developed at the cathode of the blanking gate stage, V18, through the conduction of CR5. Because of the increase in bias voltage, this stage is held cutoff for the duration of the pulse at its cathode. The blanking procedure is needed to eliminate undesirable feedback from the klystron to the receiver.

FREQUENCY MULTIPLIER—OSCILLATOR

The frequency multiplier-oscillator (fig. 15-11) contains the oscillator chassis, the pulse shaper chassis, and three coaxial amplifiers used to multiply the oscillator output frequency. The oscillator, V1, produces the fundamental operating frequency for use in the circuits of the transmitter, and the basic local oscillator frequency for the receiver. The operation of the crystal oscillator (fig. 15-4) is considered in the discussion of the local oscillator circuits in the receiver.

BUFFER AND MULTIPLIER STAGES

V1 comprises the oscillator stage, which is crystal controlled by Y1. The output of the oscillator (between 40 and 50.5 mc) is tuned by C8 and coupled from a tap on the T2 secondary via C9 to the grid of the buffer stage, V2.

The V2 grid draws a small current for a part of each cycle, and grid leak bias is developed across R8. Cathode bias is developed at the V2 cathode by R9. When the oscillator is properly tuned, the grid current of the buffer stage will be small. Thus, the input impedance to V2 is relatively high so that the effects of keying in the grid circuit of the first doubler, V3 (as discussed) will not cause serious reflections back into the oscillator through large values of buffer grid current.

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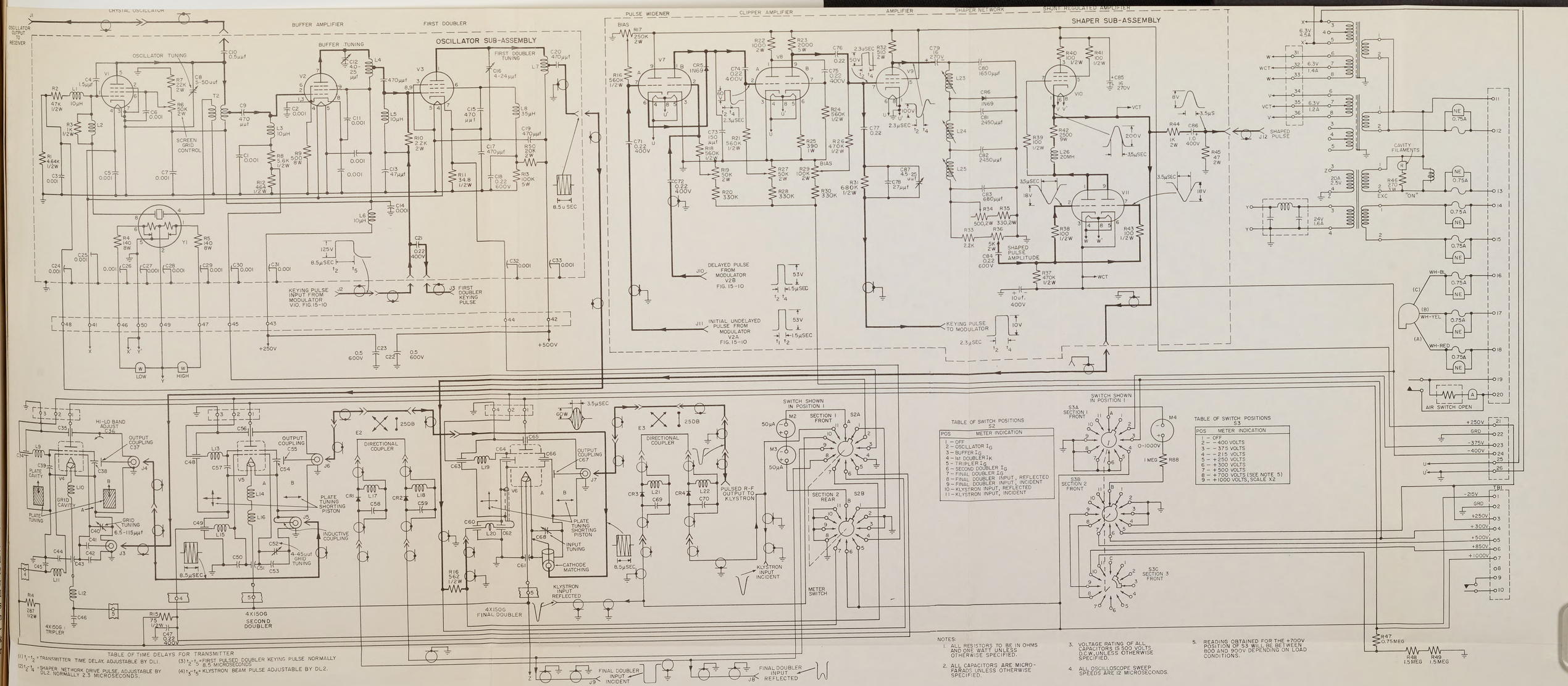
OSCILLATOR SUB-ASSEMBLY
SHAPER SUB-ASSEMBLY
CRYSTAL OSCILLATOR
BUFFER AMPLIFIER
FIRST DOUBLER
PULSE WIDENER
CLIPPER AMPLIFIER
AMPLIFIER
SHAPER NETWORK
SHUNT REGULATED AMPLIFIER
OSCILLATOR OUTPUT TO RECEIVER
OSCILLATOR TUNING
BUFFER TUNING
FIRST DOUBLER TUNING
SCREEN GRID CONTROL
KEYING PULSE INPUT FROM MODULATOR V10. FIG. 15-10
FIRST DOUBLER KEYING PULSE
DELAYED PULSE FROM MODULATOR V2B FIG. 15-10
INITIAL UNDELAYED PULSE FROM MODULATOR V2A FIG. 15-10
KEYING PULSE TO MODULATOR
SHAPED PULSE AMPLITUDE
SHAPED PULSE
CAVITY FILAMENTS
EXC
"ON"
AIR SWITCH OPEN
LOW
HIGH
+250V
+500V
WCT
VCT
HI-LO BAND ADJUST
OUTPUT COUPLING
PLATE CAVITY
GRID CAVITY
PLATE TUNING
GRID TUNING
INDUCTIVE COUPLING
PLATE TUNING SHORTING PISTON
DIRECTIONAL COUPLER
PULSED R-F OUTPUT TO KLYSTRON
INPUT TUNING
CATHODE MATCHING
KLYSTRON INPUT REFLECTED
KLYSTRON INPUT INCIDENT
METER SWITCH
4X150G TRIPLER
4X150G SECOND DOUBLER
4X150G FINAL DOUBLER
FINAL DOUBLER INPUT INCIDENT
FINAL DOUBLER INPUT REFLECTED
SWITCH SHOWN IN POSITION 1
SECTION 1 FRONT
SECTION 2 REAR
SECTION 2 FRONT
SECTION 3 FRONT
TABLE OF SWITCH POSITIONS S2
POS METER INDICATION
1 – OSC
2 – OSCILLATOR I_G
3 – BUFFER I_G
4 – 1st DOUBLER I_K
5 – TRIPLER I_G
6 – SECOND DOUBLER I_G
7 – FINAL DOUBLER I_G
8 – FINAL DOUBLER INPUT, REFLECTED
9 – FINAL DOUBLER INPUT, INCIDENT
10 – KLYSTRON INPUT, REFLECTED
11 – KLYSTRON INPUT, INCIDENT
TABLE OF SWITCH POSITIONS S3
POS METER INDICATION
1 – OFF
2 – – 400 VOLTS
3 – – 375 VOLTS
4 – – 215 VOLTS
5 – +250 VOLTS
6 – +300 VOLTS
7 – +500 VOLTS
8 – +700 VOLTS (SEE NOTE 5)
9 – +1000 VOLTS, SCALE X2
TABLE OF TIME DELAYS FOR TRANSMITTER
(1) t_1–t_2 = TRANSMITTER TIME DELAY, ADJUSTABLE BY DL1.
(2) t_2–t_4 = SHAPER NETWORK DRIVE PULSE, ADJUSTABLE BY DL2. NORMALLY 2.3 MICROSECONDS.
(3) t_2–t_5 = FIRST PULSED DOUBLER KEYING PULSE NORMALLY 8.5 MICROSECONDS.
(4) t_3–t_5 = KLYSTRON BEAM PULSE ADJUSTABLE BY DL2.
NOTES:
1. ALL RESISTORS TO BE IN OHMS AND ONE WATT UNLESS OTHERWISE SPECIFIED.
2. ALL CAPACITORS ARE MICROFARADS UNLESS OTHERWISE SPECIFIED.
3. VOLTAGE RATING OF ALL CAPACITORS IS 500 VOLTS D.C.W. UNLESS OTHERWISE SPECIFIED.
4. ALL OSCILLOSCOPE SWEEP SPEEDS ARE 12 MICROSECONDS.
5. READING OBTAINED FOR THE +700V POSITION OF S3 WILL BE BETWEEN 800 AND 900V DEPENDING ON LOAD CONDITIONS.

Figure 15-11.—Frequency multiplier-oscillator, schematic diagram.

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A -215-volt bias is applied to V5 through L15 and R15. The positive peaks of the signals from V4 drive the grid into the area of grid current. Grid current metering is provided when the meter switch is in the 2nd DOUBLER Ig position (position 6). R15 acts as the meter shunt, while C47 filters the meter current to eliminate the r-f component.

Screen voltage is applied to V5 through L13. Built-in capacitor C48 filters out the r-f component that enters the screen voltage line from the tube. The plate voltage is directly applied to V5, and is filtered by C56. The V5 output is coupled by C55 to coaxial connector, J6. From this point, the 480- to 612-mc signal is passed along a coaxial line to the final doubler stage, V6.

A directional coupler, E2, is connected between the second doubler, V5, and the final doubler, V6. A directional coupler is a device used to sample the power in the incident and/or reflected wave on a transmission line. The device monitors the power that moves in one direction only along the line. The power extracted for monitoring is a small portion of the power in the line; however, the magnitude of the monitoring power is proportional to the magnitude of the power propagated along the line. The coupling units are so arranged that reflected signal power does not affect the accuracy of the incident power measurements.

The incident signals from the directional coupler are developed across C59 and L18 and rectified by CR2. The rectified signal current is applied to the meters, M2 and M3, when the meter switch, S2, is placed in position 9 (FINAL DOUBLER INPUT, INCIDENT position). The reflected signal is developed across C58 and L17 and rectified by CR1. This signal is applied to the meters when the meter switch is in the FINAL DOUBLER INPUT, REFLECTED position (position 8).

The final doubler, V6, is driven by feeding the signal to a variable tap on the cathode line. The setting of this tap is determined by a minimum amplitude of the reflected signal as seen on an oscilloscopic display. The oscilloscope is connected to the final doubler input reflected jack J8, which is located on the front panel of the frequency multiplier oscillator chassis.

A 3.5-μs pulse from V10, whose leading edge is coincident in time with the arrival of a pulse from V20 (fig. 15-8) in the coder-indicator is applied at the grid of V6 through R16 and L20 and gates the V6 final doubler stage. In the absence of this pulse, the negative potential from the cathode of V10 is sufficient to hold V6 beyond cutoff. The origin of the 3.5-μs pulse is considered later in this chapter.

The V6 grid cavity is tuned by C68 to the 480- to 612-mc input from the second doubler. This cavity is tuned by pushing the capacitor plunger in or out. The plate cavity is tuned by a shorting piston. This cavity has an equivalent electrical length of three-quarters the wavelength corresponding to the output frequency (960 to 1214 mc).

The output of the final doubler, V6, is coupled by C67 to the final doubler output jack, J7. This signal is then applied to a directional coupler, E3. The operation of this circuit is the same as that of the directional coupler, E2, described earlier, and is used to measure the respective amplitudes of the incident and reflected components of the signal fed to the klystron. The incident component of the E3 output is measured when the meter switch, S2, is in the KLYSTRON INPUT, INCIDENT position (position 11). When the switch is in position 10, the reflected component of the klystron input can be observed.

PULSE WIDENER AND PULSE SHAPER STAGES

The shaper network in the frequency multiplier-oscillator receives rectangular pulses from cathode follower V2 in the low-level modulator (fig. 15-10). The shaper chassis converts, or shapes, the input pulses so that they have the proper amplitude and duration for keying the final doubler circuit, V6.

Full traffic for the transmitter consists of the reference pulse groups, the identification pulses, and the distance reply pulse-pairs. However, this discussion of the shaper chassis concerns itself with the generation of one shaped pulse.

For each pulse entering the amplifier-modulator chassis (fig. 15-10) from the coder-indicator (fig. 15-8), two pulses from V2 (fig. 15-10) the low-level modulator enter the pulse widening circuit of the shaper chassis in the frequency multiplier-oscillator through J10 and J11 (fig. 15-11). One pulse is delayed from the other by an amount determined by the taps on the delay line, DL2, in the low-level modulator (fig. 15-10). Each of the pulses is approximately 1.5 μs wide at the half-amplitude points.

The delay between the two pulses can be varied from 1.8 to 2.7 μs in steps of 0.15 μs. The nominal value of delay is 2.3 μs.

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The pulse widener circuit, V7, of the shaper chassis functions essentially the same as the pulse widener circuit considered in the discussion of the low-level modulator (V2, fig. 15-10). V7A is biased beyond cutoff since the cathode is at a sufficiently higher positive potential than the grid. V7B is also biased beyond cutoff by a negative voltage from the -375-volt supply. This potential is applied from the arm of R19, which is a portion of a voltage divider comprising R19 and R20. The divider is connected from the -375-volt supply to ground.

Each of the positive pulses applied to V7 has an amplitude of approximately 53 volts. The initial pulse causes V7A to conduct through CR5 to further increase the charge on C73 for a total duration of 1.5 μs. Until such time as V7B conducts, the only partial discharge path for C73 is through C74 and the relatively high resistance of R21.

The time of arrival of the delayed pulse to the V7B grid is approximately 2.3 μs after the beginning of the leading edge of the undelayed pulse at V7A. When the delayed pulse is applied to V7B, this tube section conducts, providing a low resistance discharge path for C73. Because of this arrangement, the voltage at the CR5-C73 junction is held at approximately 60 volts (with respect to ground) for a total period of 2.3 μs.

The output pulse of the pulse widener circuit is applied through C74 to the control grid of the clipper-amplifier, V8A. Bias is applied to the grid of V8A from the arm of R27. This resistor forms a portion of a voltage divider comprising R27 and R28, which is connected from the -375-volt supply to ground.

Because of the biasing arrangement just discussed, the initial portion of the pulse from the pulse widener circuit will not be of sufficient amplitude to cause V8A conduction, and the baseline of the pulse is clipped. The negative-going output at the plate of V8A is coupled by C75 to the control grid of V8B. This stage is biased by current flow through the cathode resistor, R25, which keeps the value of plate dissipation within the rated value of the tube.

The negative input from V8A cuts off the V8B section of the tube, causing the V8B plate voltage to rise. C76 charges with a corresponding rise in voltage. (The total voltage of C76 is the sum of the voltage at the arm of R29 and the V8B plate voltage.) The rise in voltage (approximately 50 volts having a duration of 2.3 μs) is applied to the grid of amplifier, V9. This pulse is also applied to the shunt-regulated amplifier comprising V9, V10, and V11 in the amplifier modulator chassis (fig. 15-10), and is used in producing the 8.5-μs output. C77 (fig. 15-11) couples this signal to V9 (fig. 15-10), while C78 and C87 form a frequency compensating network, which determines the sharpness of the leading edge of the 2.3-μs pulse.

The plate load of V9 is the parallel combination of R32 and the shaper network. C79 is the coupling capacitor for the V9 output and a d-c blocking capacitor to keep the d-c voltage from appearing on the network.

The shaper network is a low-pass filter, which has a characteristic impedance of approximately 550 ohms. The components of the shaper network are L23, L24, L25, C80, C81, C82, C83, R34, and R35. When the 2.3-μs pulse is fed into this network, a symmetrical output pulse of increased duration and rise time is produced. CR6 conducts if the output of V9 moves in the positive direction to eliminate this portion of the pulse. Potentiometer R34, in series with R35, is used to match the output of the filter.

R33 and potentiometer R36 form a voltage divider network, allowing the potential at the arm of R36 to be fed to the shunt-regulated amplifier comprising V10 and V11. This control represents the only method of varying the output amplitude of the shunt-regulated amplifier.

The shunt-regulated amplifier consists of two tubes, V10 and V11, which are connected between the -400-volt bias line and the +250-volt B+ power supply. The two halves of V11 are paralleled. During quiescence this tube has a conduction of 34 ma. The negative output pulses of the shaper network are coupled by C84 to the control grid of V11. Each of these pulses causes a large decrease in the tube conduction. This action is equivalent to an increase in the effective impedance of V11, causing the current flow and voltage drop across R42 to decrease. As the voltage across R42 decreases, the bias applied to V10 also decreases, and the V10 current increases. The path for this transient current is through R45, C86, V10 and the B supply.

Since the effective impedance of V11 has increased, a large transient discharging current passes through C86. This capacitor is a part of a voltage divider comprising R44, C86, and R45, which is connected from the cathode of V10 to ground. As C86 discharges, the cathode voltage goes less negative, developing a 200-volt peak

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pulse having a duration of 3.5 μs. This pulse is used to key the final doubler, V6, as previously discussed.

That portion of the shunt-regulated output developed across R45 is made available at shaper output jack J12. This jack is located on the front panel of the frequency multiplier-oscillator chassis.

A voltmeter, M4, is used to measure the d-c voltage output of the various d-c supplies. The meter is mounted on the front panel of the frequency multiplier-oscillator chassis. The value of voltage read by the meter is determined by the setting of S3, which connects M4 to the circuit to be measured. Both the meter and multiplier resistor, R88, are placed in the circuit by S3.

RELATIVE TIME OF WAVESHAPES IN RADIO BEACON TRANSMITTER

Figure 15-12 shows a composite picture of the waveshapes developed in the radio beacon transmitter. The 2.3-μs output of V7 in the shaper chassis (fig. 15-11) is initiated by the undelayed output of V1 in the amplifier-modulator chassis (fig. 15-10). The trailing edge of the 2.3-μs pulse corresponds in time with the leading edge of the delayed pulse from the delay line, DL2 (fig. 15-10). This pulse is coupled to a shaper network, which acts upon the pulse to produce an output whose width at midamplitude is approximately 3.5 μs. The pulse has a slow rise as shown and is used to key the final doubler stage, V6, of the frequency multiplier-oscillator (fig. 15-11). Due to the time constants in the final doubler stage, this pulse is delayed slightly.

In order for the klystron, V15 (fig. 15-10) to allow the passage of the radio frequency from the final doubler (duration 3.5 μs), the klystron beam pulse must be present at the cathode. The beginning of the klystron beam pulse corresponds in time with the leading edge of the pulse to the pulse widener circuit, V3A, of the amplifier-modulator (fig. 15-10). This pulse is delayed 1.5 μs by DL2. The output pulse of the pulse widener circuit is held at 7 μs by the arrival of a triggering pulse to V3B from the delay line, causing the trailing edge of the pulse to occur at this time.

Because of the slow rise of the shaper output pulse, the portion of the pulse that triggers the final doubler to produce the r-f output to the klystron appears at approximately the center portion of the klystron beam pulse.

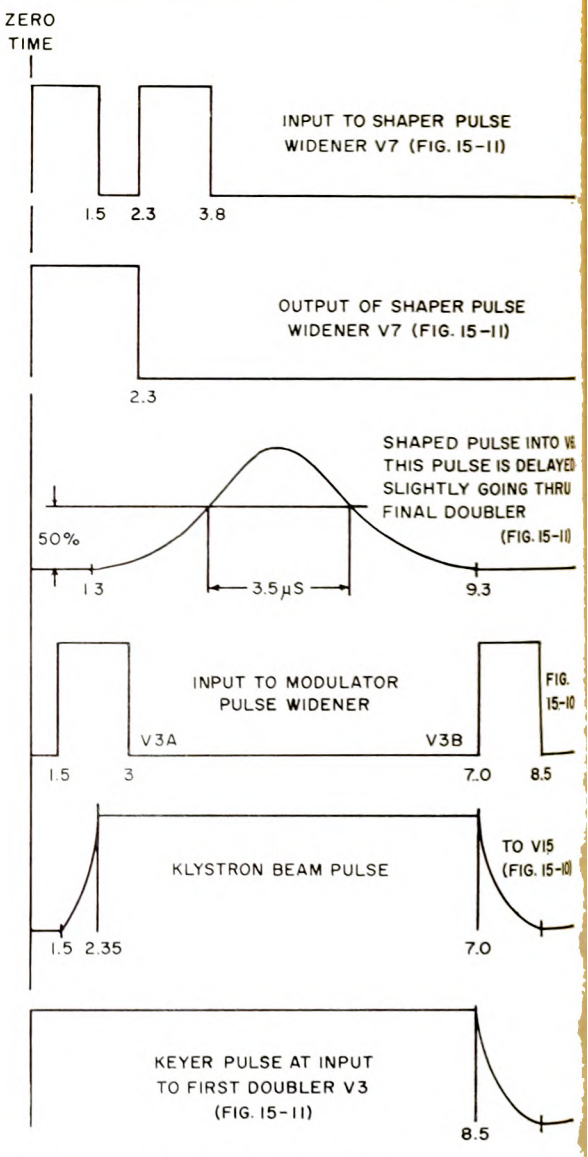

Figure 15-12.—Radio beacon transmitter, relative times of waveshapes.

The pulsed r-f power input from the final doubler, V6 (fig. 15-11) to the klystron is between 30 and 80 watts. The beam power pulse has an amplitude of 500 volts. As the two pulses appear in coincidence at the klystron, the output power of the klystron is about 5.7 kw.

The time of the pulse to key the first doubler of the frequency multiplier-oscillator corresponds with the beginning of the 2.3-μs pulse (zero time). This pulse along with the 7-μs pulse at the grid of V11 in the amplifier-modulator chassis is used to produce the 8.5-μs keying pulse for the first doubler stage.

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QUIZ

1. A filter network that allows both the transmitter and receiver to be operated simultaneously from the same antenna is defined as what type filter?
2. The interrogation pulse-pair carrier frequency will be between what limits for low and high band reception?
3. The circuit of the oscillator (fig. 15-4) is that of a crystal-controlled oscillator with what type feedback?
4. What is the frequency range of the oscillator output (fig. 15-4)?
5. The frequency multiplier section of the radio receiver (fig. 15-5) multiplies the oscillator output within what limits?
6. What is the major reason for operating the receiver and transmitter from the same fundamental frequency source?
7. The output tuned circuit of the tripler stage (fig. 15-5) is adjustable over what frequency range?
8. The input to the quadrupler stage, V5 (fig. 15-5) is taken from a tap on L11 for what reason?
9. All changes in the tacan system final output frequency that are greater than what value necessitate the retuning of the frequency multiplier stages (fig. 15-5)?
10. Echo suppression in the receiver is obtained by means of RC discharge networks in the grid circuits of what stages (fig. 15-5)?
11. Squitter control in the form of a negative voltage is fed to the grid of what stages in the receiver (fig. 15-5)?
12. The video chassis decodes the interrogation pulse-pairs and a sufficient number of noise pulses to maintain what number of pps at the receiver output?
13. When the first pulse of an interrogation pulse-pair, or the first positive pulse of a pair of noise pulses separated by 12 μs arrives at the suppressor grid of V20 (fig. 15-5), it is not applied simultaneously to the control grid of the tube because of what component?
14. The characteristic of the decoder (V20, fig. 15-5) insures that only those pulses (noise or interrogation)having what spacing will be acted upon the the decoder circuit?
15. The decoder output consists of how many pulses for each decodable pulse-pair input?
16. The blanking circuit (fig. 15-5) is responsible for the blocking of the decoder input for what duration after a pulse-pair has entered the circuit?
17. The output of the V21B multivibrator cathode (fig. 15-5) is applied via C125 to the control grid of the pulse amplifier, V26, and is used in deriving what voltage?
18. Since without the application of bias to the grid of V24 the voltage drop across R107 may result in a high positive voltage (in excess of +20 volts, d-c) applied to the i-f amplifier, what tube operates to prevent the voltage from rising above +5 volts while the receiver is warming up?
19. The coder-indicator (fig. 15-7) receives its input from what two sources within the radio beacon?
20. What is the frequency output of V22 (fig. 15-8) relative to the output frequency of the oscillators?
21. In the 135-cps reference burst circuit (fig. 15-5) after six pulses have been delivered by the frequency dividers, the action must be terminated by closing the gate being applied to V5 from what stage?
22. The 135- and 15-cps gates coupled from the cathode of the respective 135- and 15-cps pulse gate multivibrators (fig. 15-8) are applied to V15 for the purpose of blanking the distance reply signals or the identification call signals during what period?
23. The 1350-cps oscillator, V11, receives the negative 132-μs gate from the 135-cps pulse gate multivibrator (fig. 15-8), which is used for what purpose?
24. For each pulse received from the double coding tube, V18 (fig. 15-8) the output multivibrator produces how many output pulses?
25. The input to the amplifier-modulator (fig. 15-10) arriving from the coder-indicator consists of how many pulses?
26. The width of the pulse developed at C10 (in the V3B plate of figure 15-10) is determined by what factors?
27. The circuit of the high-level modulator (fig. 15-10) consists of a series arrangement of what stages?
28. What tube acts as the a-c load on the high-level modulator (fig. 15-10)?
29. What are the three general circuits comprised in the frequency multiplier-oscillator (fig. 15-11)?
30. What is the device used to sample the power in the incident and/or reflected waves on the transmission line between stages V5 and V6 (fig. 15-11)?
31. In order for the klystron, V15 (fig. 15-10) to allow the passage of the radio frequency from the final doubler (duration 3.5 μs), what other conditioning pulse must be present?

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CHAPTER 16

ELECTRONIC AIDS TO NAVIGATION: TACAN (PART III)

INTRODUCTION

The theory of operation of the antenna control circuits, including the antenna azimuth (bearing) circuits, the antenna roll and pitch circuits (used at shipboard installations) and the antenna speed control circuits, is considered in this chapter. A block diagram approach to the circuits of the antenna system was presented in chapter 14 of this training course. It is necessary to read that material before an attempt is made to study the contents of this chapter.

ANTENNA GROUP

The antenna group (chapter 14, figure 14-1) consists of the antenna, the antenna base, and the antenna control system. A magnetic variation subassembly (chapter 14, figure 14-17) is a part of the antenna bearing (azimuth) servo control system, and is therefore considered in the discussion of the antenna group.

The antenna group of the radio beacon (1) receives distance interrogation pulse-pairs from the aircraft, (2) radiates replies (which the aircraft receive and processes to obtain distance and bearing information), (3) provides bearing reference pulses to trigger the coder-indicator at specific intervals, (4) modulates the r-f pulse envelope with the 15- and 135-cps frequencies used for bearing information, (5) stabilizes the antenna (in the case of shipboard installations), and (6) provides for the automatic control of the speed of rotation of the antenna.

There are two possible antenna groups for use with the shore radio beacon, and four possible antenna groups for use with shipboard installations. Each one of these antenna groups consists of several nomenclatured components, including the antenna proper, the antenna base, and the antenna control unit.

REFERENCE PULSE GENERATOR

The central (stationary) antenna array (see figure 14-4, A, chapter 14) contains the primary radiating elements, consisting of seven biconical dipoles mounted in a fiberglass cylinder. The 15-cps modulation is accomplished by parasitic reflectors mounted in a separate fiberglas cylinder, which rotates around the central array at a constant speed of 900 rpm. The antenna used for low-band transmission and reception has its modulation introduced by two resistance wires (similar to that shown) spaced close together in the cylinder. The high band antenna has four such wires with a similar close spacing to produce the modulation. The 135-cps modulation results from a second cylinder, mounted outside the first, which is rotated from the same shift that drives the 15-cps cylinder. The 135-cps cylinder contains eight parasitic reflectors, which are spaced evenly around the cylinder.

The rotating shaft that moves the parasitic elements around the central array has a pulser plate (attached to it), into which is embedded nine soft-iron slugs, as discussed in chapter 14. These slugs, passing through a pulser coil, generate pulses, which, in turn, trigger the coder-indicator unit of the radio beacon. This action was described in chapter 15. The pulser plate, in combination with the pulser coil assembly, is known as the reference pulse generator.

The precise sequence and timing of the trigger pulses generated by the reference pulse generator are determined by the spacing of the soft-iron slugs about the outer edge of the pulser plate. As stated earlier, the reference pulse generator provides two sets of reference pulses, corresponding to the 15- and 135-cycle components of the bearing signal. These 15- and 135-cps reference pulses trigger the coder-indicator periodically.

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The 15-cps pulse triggers the coder-indicator, 15-cps circuit once in every 360° rotation of the antenna. The 135-cps pulses trigger the 135-cps circuit of the coder-indicator every 40° of antenna rotation after the 15-cps trigger pulse. Figure 14-5, B (chapter 14) shows that the eight 135-cps slugs are spaced 40° apart about the lower edge of the pulser plate with the exception of the 80° blank space. The north 15-cps slug triggers the coder-indicator at this time.

Note that the north slug is placed 180° from the blank space. This is done so that the pulser coil for the north reference trigger may be placed on the opposite side of the pulser plate. Thus, during an 80° interval, the 135-cps pulser coil is not activated by a slug on the 135-cps cylinder. However, during this same period, the slug on the 15-cps cylinder is active in supplying the 15-cps trigger to its circuits in the coder-indicator.

The reference signals of the 15- and 135-cps trigger pulses fed to the coder-indicator are processed and fed, in turn, to the transmitter. The transmitter circuits shape the signals and return them to the antenna array as r-f pulses for transmission.

GENERAL THEORY OF SERVO AND SYNCHRO SYSTEMS

To facilitate an understanding of the shipboard and shore antenna control systems, a brief discussion on general servo systems is presented in the following paragraphs. For a more detailed treatment of this subject, refer to Chapter 17, *Basic Electricity*, NavPers 10086.

The transmission of the angular position of a mechanical shaft to a remote location may be accomplished, electrically, by a synchro system. If the voltage representing the angular position of the controlling shaft is to be amplified before it is applied to the controlled shaft, a servo system is used instead. Servo systems are essentially high-gain power amplifiers operating on the closed loop or error-sensitive principle in which the action of the power amplifier is governed by an error that is developed when there is an angular difference between the controlling and controlled shafts.

The controlling system may employ a synchro (transmitter and receiver), a differential transmitter, or a control transformer. The system may also employ any combination of these units to accomplish the controlling action.

The function of a synchro transmitter is to transmit electrical data corresponding to the angular position of its rotor. A synchro transmitter consists of a single-winding, 2-pole rotor and a wye-connected three-winding, 2-pole stator (fig. 16-1). The stator is wound so that the three windings are displaced by 120° about the axis of the rotor shaft. When an a-c voltage is applied to the rotor, electrical signals induced in the stator windings are proportional to the angular position of the rotor.

The electrical characteristics of a synchro receiver are identical to those of a synchro transmitter just discussed. Physically, however,

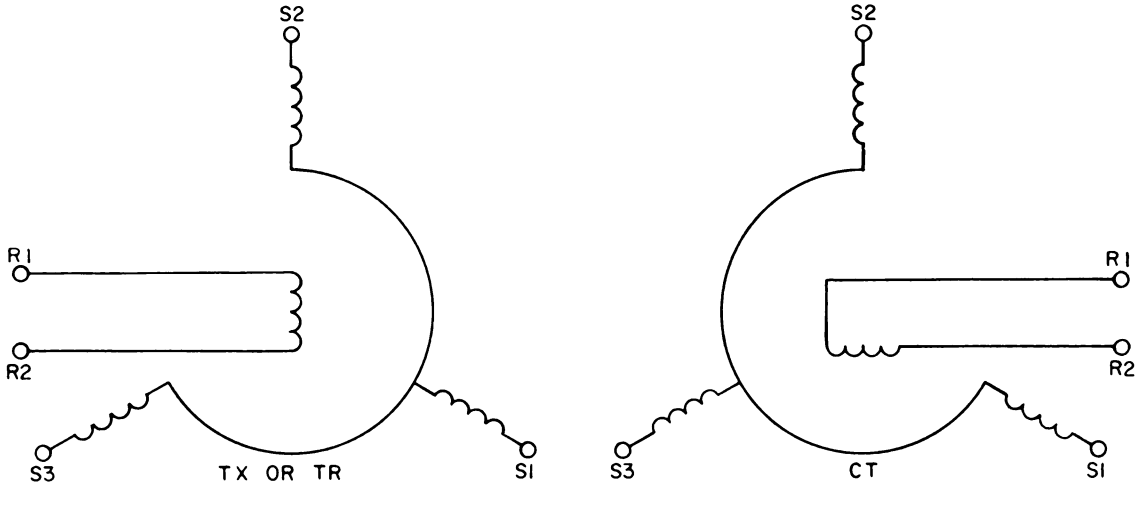

Figure 16-1.—Synchro transmitter (or receiver) and synchro control transformer.

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the receiver differs from the transmitter in that it is equipped with an inertial damper and has very low friction losses. The inertial damper prevents the rotor from overshooting caused by any sudden changes in the received data. If a sudden change in the incoming voltage causes the receiver to alter its position abruptly, a damping ring exerts a damping or braking action on the oscillating rotor, which stops its movement. When the change in incoming voltage results in a steady movement, the ring moves with the shaft, and no damping action occurs. The synchro receiver is normally connected in parallel with the synchro transmitter to provide an indication of the angular position of the transmitter rotor at the remote location.

A synchro control transformer is similar to a synchro transmitter and synchro receiver except for some physical and electrical differences. The a-c supply voltage is not applied to the rotor winding of the control transformer. Instead, a voltage is induced in the rotor winding by the current flow in the stator windings. The control voltage reaches a null (zero volts output) when the control transformer rotor is displaced 90° from the rotor of the transmitter synchro. When the rotor of a transmitter synchro is displaced (and therefore out of alignment with the control transformer rotor) the voltage that appears across the control transformer rotor winding has an amplitude that varies as the sine of the angle of displacement.

The control transformer is used where it is desired to obtain a voltage indicative of the error or difference in the angular positions of the input and output shafts of a servomechanism. This voltage is then fed to a servo amplifier, which, in turn, drives a motor geared so as to restore the alignment of the shaft of the control transformer to the null position.

The phase of the error signal from the control transformer depends on the direction in which the rotor of the control transformer is out of correspondence with the rotor of the transmitter. The output voltage, or error signal, usually controls magnetic amplifiers, which feed power to servo motors. The motors are mechanically geared to the shaft of the control transformers. Thus, the motor rotates until the rotor of the control transformer is restored to the null position.

The differential transmitter (fig. 16-2) provides a means of adjusting the angular position of the receiver rotor with respect to the transmitter rotor. The stator of the differential transmitter is similar to that of the regular synchro transmitter or receiver. However, the differential transmitter rotor differs physically and electrically from the rotor of conventional synchro units.

The synchro differential transmitter may be employed between a synchro transmitter and a control transformer, or between a synchro transmitter and a synchro receiver. Figure 16-2, A shows a synchro differential transmitter connected between a synchro transmitter and a synchro control transformer.

The stator of the differential transmitter represents the primary, while the rotor represents the secondary of a variable transformer. The transformer ratio is approximately 1:1.

In the arrangement shown, the signal from the synchro transmitter is fed to the stator of the differential transmitter, and its output signal is fed to the stator windings of a control transformer. The variable introduced in such a control system by the differential transmitter is a function of the angle of the differential transmitter rotor position, which can add or subtract from the angle introduced by the synchro transmitter.

To explore the action of this circuit arrangement, assume that the differential transmitter rotor (fig. 16-2, A) is at the electrical zero position when the transmitter rotor is turned counterclockwise to the 60° position. Since the differential transmitter is at zero, no correction factor is added, and the control transformer is aligned with the transmitter rotor when the system is in this stable or null position. On the other hand, if the differential rotor is turned counterclockwise to the 120° position with the transmitter rotor set at 0 degrees (fig. 16-2,B), the control transformer will be turned clockwise by the servo system to the 240° or null position.

When the differential generator is directly connected between the transmitter and receiver, the null position of the control transformer shaft will be equal to the position of the transmitter shaft minus the position of the differential transmitter shaft.

By reversing the S1-S3 and R1-R3 connections of the differential generator, the differential transmitter shaft setting will add to the synchro transmitter reading. The magnetic variation subassembly (discussed later) used at shore installations employs a differential transmitter, which is connected as a synchro transmitter.

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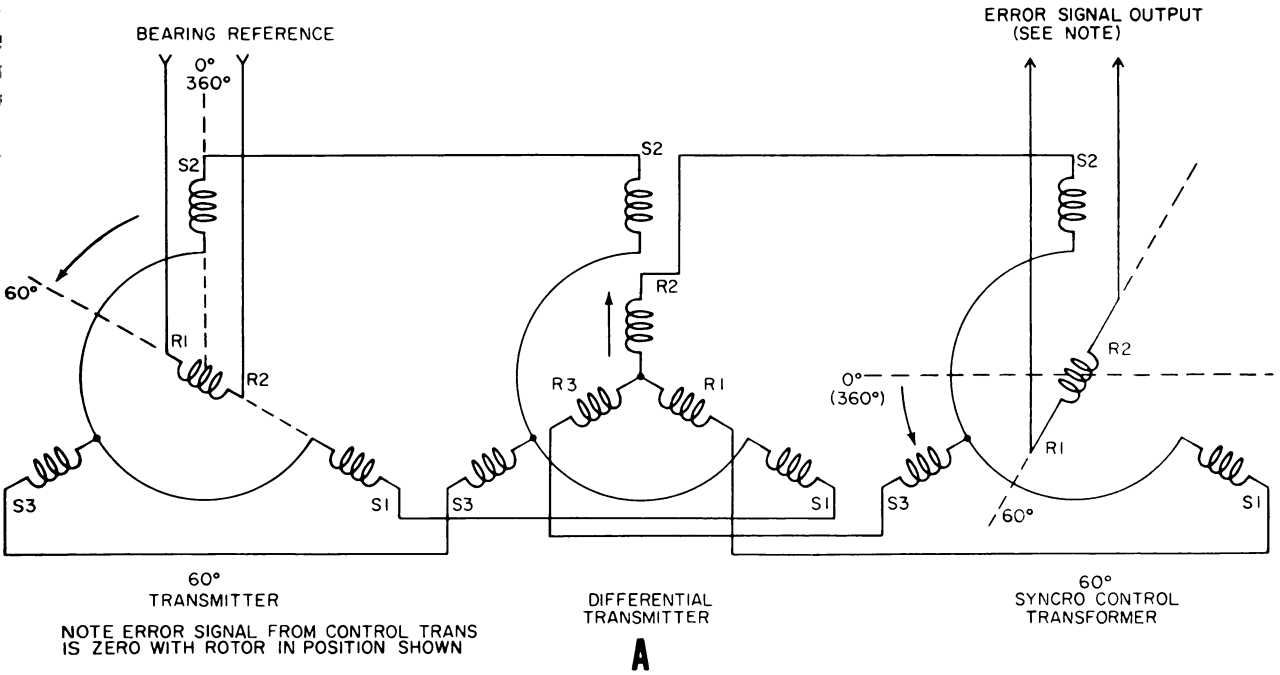

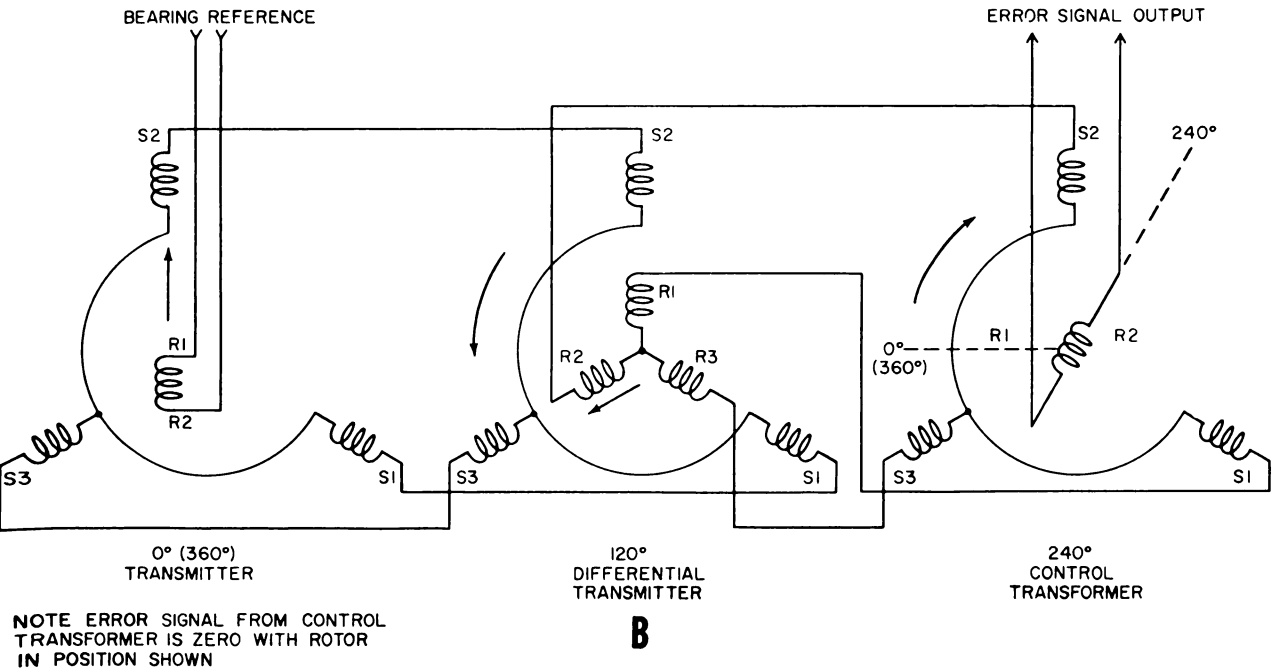

Figure 16-2.—A. Differential transmitter at 0° shown connected to a synchro transmitter at 300°.
B. Differential transmitter at 240° to set control transformer at 120°.

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MAGNETIC VARIATION SUBASSEMBLY

The angular position signals for true north transmitted from the ship's compass are continuously fed through the ship's bus to the magnetic variation subassembly of the receiver-transmitter group (see figure 14-17, chapter 14). The signal maintains the radio beacon's 15-cps reference pulse coil subassembly fixed with respect to corrected (magnetic) north.

The magnetic variation signals from the gryo compass (in shipboard equipments) are fed to the 1- and 36-speed differential transmitters, corresponding to DG1 and DG2, respectively of figure 16-3. C1 and C2 are power factor correction capacitors for the voltage applied to the rotors of DG1 and DG2.

The input signals to the differential transmitters (at the stators) cause a voltage in the secondary rotor, which is applied to the 1- and 36-speed control transformers, CT1 and CT2, respectively. C3 and C4 are power factor correction capacitors for the input circuit to the control transformers. The output of the control transformers is the error voltage, which is applied to the bearing magnetic amplifier circuit through the closed contacts 6-5 and 4-3, respectively, of K1.

The magnetic correction signals may be manually set into the 1- and 36-speed differential transmitters from the front panel of the magnetic variation subassembly (see chapter 14, figure 14-17). The sum of the gyro compass signals and the magnetic variation signals is forwarded to the azimuth 1- and 36-speed control transformers (fig. 16-3), as discussed. The control transformers are located in the antenna base.

As long as the rotors of the control transformers are at the same effective angular position as that of the differential transmitters (after the summing-up of the ship's gyro compass and magnetic variation signals), no control transformer output is sent to the bearing magnetic amplifier. If the equilibrium is upset due to the displacement of the rotors of the differential transmitters to new positions (by changing the setting of the magnetic variation unit) or if the ship's course changes, an error voltage will be induced and transmitted from the control transformer to the servo bearing magnetic amplifier.

(Fundamentally, the magnetic variation servo system of the shore-based beacon functions in the same manner as described for the shipboard beacon bearing servo system, except that the angular position signal received by the control transformers is generated only by the magnetic variation subassembly.)

BEARING MAGNETIC AMPLIFIER (MAGNETIC AMPLIFIERS, INC. DESIGN)

The bearing magnetic amplifier (fig. 16-4) detects the phase of the error signal from the control transformer, and amplifies the signal to a level sufficient to drive the bearing servo motor. The motor, in turn, positions the pickup coils of the pulser plate assembly (coupled from a gear train on the motor), which enables the 15-cps reference burst of the transmitted signal to be oriented with respect to magnetic north as dictated by the ship's course and magnetic variation subassembly.

The phase of the input error signal may vary with respect to a reference voltage. To detect a reversal in phase of the a-c signal, it is necessary to compare the signal with the reference voltage. This action is accomplished through the use of the 1- and 36-speed phase sensitive detectors. The one-speed error signal from the one-speed control transformer (CT1 of figure 16-3) is applied to the one-speed phase sensitive detector (fig. 16-4) at the primary (terminals 1-2) of T1. This transformer is connected (in parallel with T4) from terminal 5 of TB1 through the secondary of T5 and returned to terminal 6 of TB1. At T5, a 5-volt antistickoff potential is fed to the secondary, which places this voltage in series with the one-speed error voltage. R19 and R20 form a voltage divider at the antistickoff voltage input, which reduces the potential applied across the transformer. C7 and R18 shift the phase of the voltage, as required. The action of the antistickoff voltage in the one-speed error input is introduced in chapter 14 and described in more detail later in this chapter.

The 117-volt, 60-cycle reference voltage required in the operation of the phase sensitive detector is obtained from the gyro amplifier via terminals LCC and LC. This voltage is fed to the 6-7 windings of T1 and T2 in parallel, to the 6-7 winding of T4 (via R14 and C6) and to the antistickoff input, T5.

The error voltage to the one-speed phase sensitive detector is obtained from the one-speed control transformer, CT1 (fig. 16-3). The 36-speed error input is simultaneously applied to the 36-speed phase detector from the 36-speed control transformer, CT2. The magnitude of the

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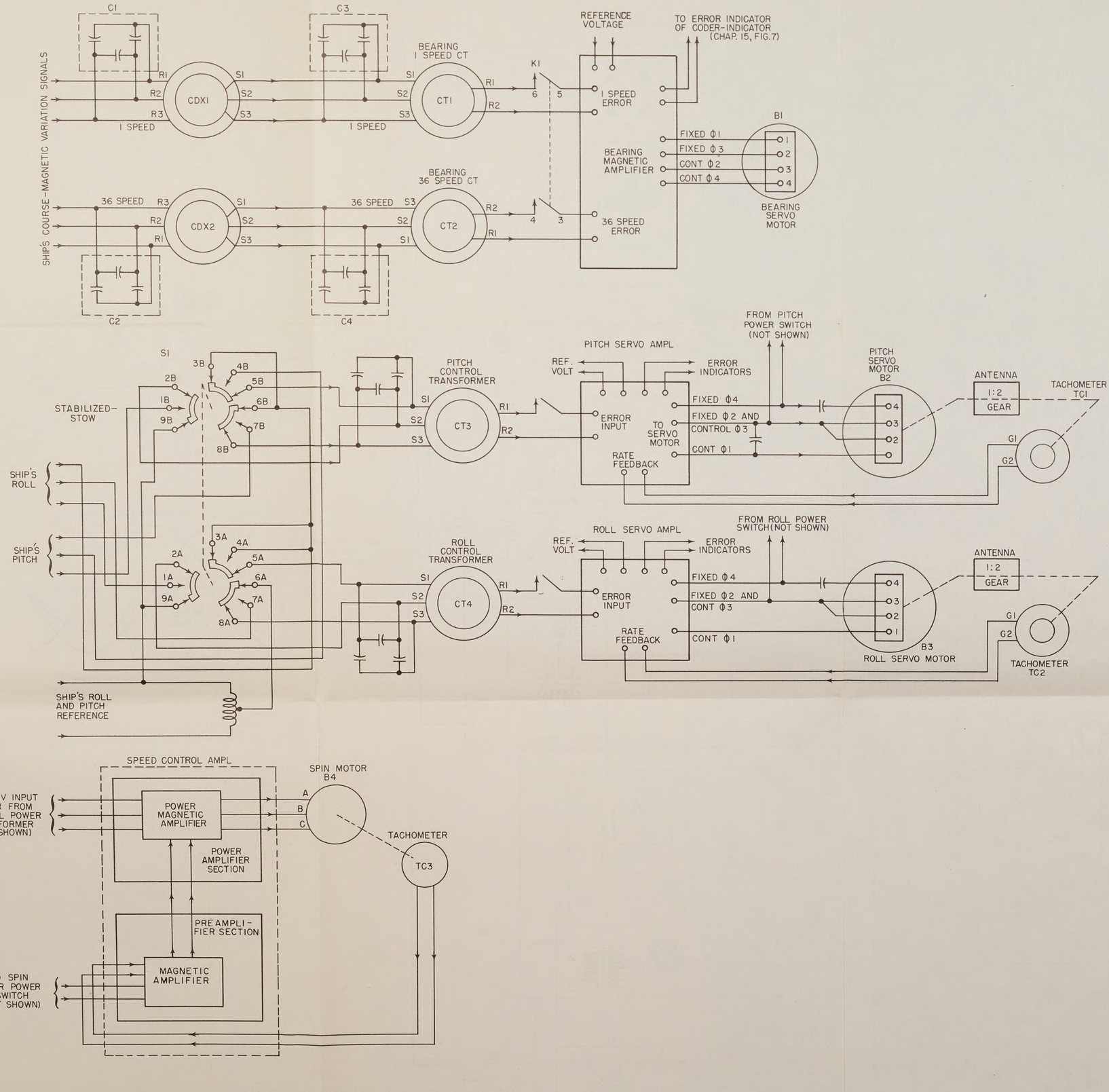

Figure 16-3.—Composite antenna group system connections, simplified diagram.

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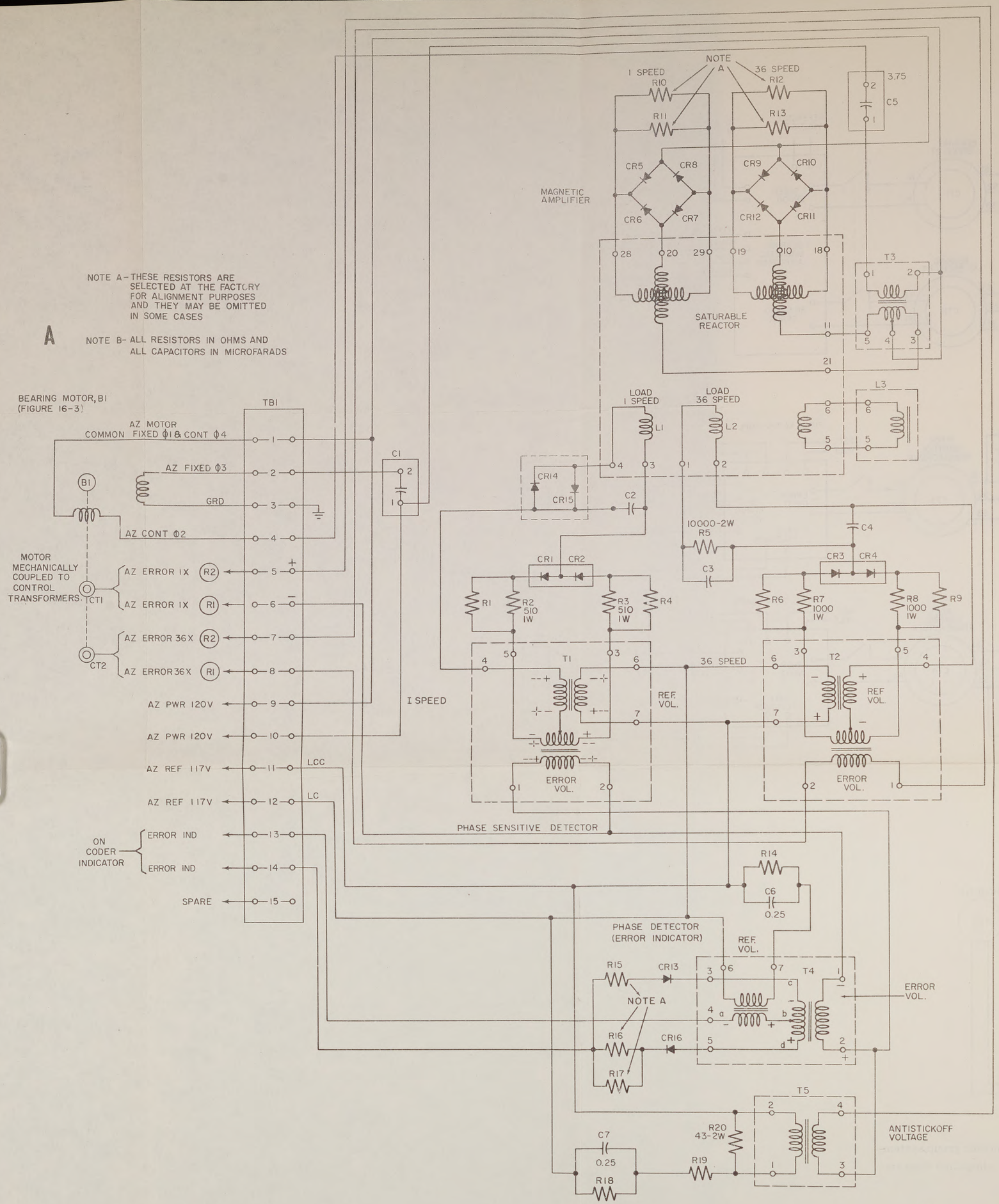
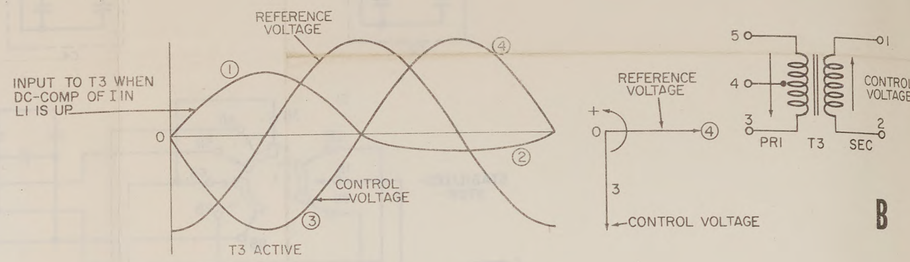
NOTE A- THESE RESISTORS ARE SELECTED AT THE FACTORY FOR ALIGNMENT PURPOSES AND THEY MAY BE OMITTED IN SOME CASES

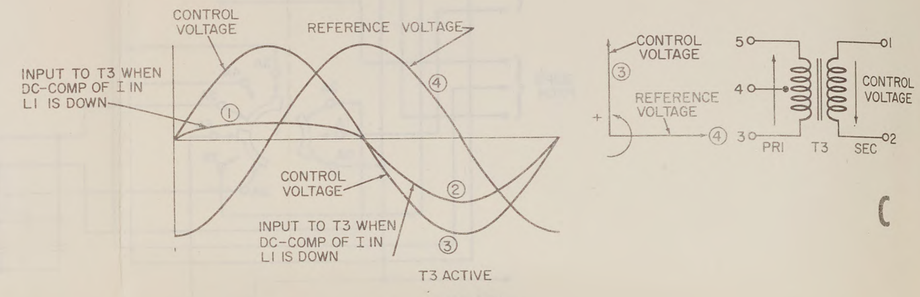
NOTE B- ALL RESISTORS IN OHMS AND ALL CAPACITORS IN MICROFARADS

A





B



C

Figure 16-4.—Bearing magnetic amplifier (Magnetic Amplifier, Inc.), schematic diagram and waveforms

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ror voltage to the 1- and 36-speed phase tectors is a function of the rate of change of e shaft position. The phase of the error is de- rmined by the direction of movement of the tor of the control transformer.

If the error signal from the differential ansmitter (fig. 16-3) exceeds 2.5 volts (the uivalent of a shaft position 2.5° away from ll) the one-speed control transformer and the e-speed phase sensitive detector will operate produce the control signal (via L1) to the e-speed magnetic amplifier. For shaft posi- ons less than 2.5° away from null (equivalent voltages below 2.5 volts) the one-speed ntrol transformer error to the one-speed ase detector will not be sufficient to cause nduction through CR14 and CR15 in the one- eed circuit. However, the 36-speed detector hich is also receiving the error) will operate produce the controlling signal, and the one- eed output is deemphasized.

To analyze the operation of this circuit, ssume that the instantaneous quiescent polari- es are as shown in solid lines at the input to e one-speed phase detector, T1 (fig. 16-4, A). lso assume that the input error is in excess 2.5°, and the one-speed detector is producing e output control current to the control phase motor B1 (fig. 16-3).

It should be observed that the voltages, db d bc, (fig. 16-4, A) for the instant shown are ries opposing, while db and ab are series ding. The sum of the latter two voltages uses a current through current limiting re- stors R1 and R2 in parallel, through CR1, (upward) and CR14, to return to d on the ansformer. Note that C2 charges during the nduction period.

On the next half cycle, all polarities are anged, as shown in dotted lines. The voltages, and bd, are still additive, but are unable to oduce a current flow through the high back sistance of CR1. Although voltages bc and bd e series opposing, the magnitude of the ference voltage, bd, is larger than the bc ltage. Thus, on this half cycle, a resultant ltage, db, causes a current through CR15, L1 ownward), CR2, R3, and R4 (in parallel), and rough terminal 3 of T1, to return to point b. nce the difference in potential (bd - bc) applied this circuit is much less than for the other lf cycles, less current flows in the downward rection through L1 during this period than was ing conducted on the previous half cycle in e upward direction. Thus, the output current in L1 has a d-c component upward. C2 charges to the average value of the d-c component of voltage across L1.

If the phase of the error input is changed, the error voltage will be shifted 180° with respect to the reference voltage. The effect is to produce a current through L1, which contains a d-c component, but of the opposite direction to that just discussed.

Al alternating (60 cps) flux exists across the 20-21 winding of the saturable reactor. The alternating current in the winding is obtained from the AZ PWR 120V (pin 9) terminal of TB1, through terminals 4-3 of T3, through the 20-21 winding, through CR7, the 29-28 winding of the saturable reactor (in parallel with R10 and R11), and CR5, to return to the other side of the power input via terminal 10 of TB1. The resultant mmf developed by L1 for a given input error aids the mmf in the 20-21 winding during one half cycle to reduce the series impedance of the winding. This action increases the magnitude of the output current through terminals 3 and 4 of T3. When the input cycle reverses, the magnetomotive forces of the L1 and 20-21 windings oppose, and the series impedance of the 20-21 winding is increased. The resulting current through the T3 primary (terminals 3 and 4) has a waveform in which alternate half cycles are suppressed (fig. 16-4, B). In this case ① is normal and ② is suppressed. The T3 secondary voltage produces an alternating current. The output of T3 is the control voltage ③, which is applied to the control phase of the bearing motor, B1. If the phase of the control voltage ③ is reversed with respect to the reference voltage ④, the direction of the motor rotation will be reversed. C5 (fig. 16-4, A) tunes out the inductive reactance introduced by the windings of the saturable reactor and the transformer.

If the polarity of the input error voltage reverses with respect to that of the reference voltage, the L1 magnetomotive force will be produced in the opposite direction. The effect on the 20-21 winding of the saturable reactor is shifted 180°, and the alternate half cycle ① (fig. 16-4, C) to the T3 primary will be suppressed. This action has the effect of reversing the phase of the output voltage ③ with respect to the reference voltage ④. Thus, the motor will turn in the opposite direction.

The reference voltage winding of the motor is excited from the supply source through phase-shift capacitor C1. The capacitor causes a shift in phase of approximately 90° with respect to

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the control winding voltage in order to produce the necessary starting torque when both windings are energized.

The action of the 36-speed phase sensitive detector is similar to that described for the 1-speed circuit, with the exception that the 36-speed circuit does not contain the de-emphasis network. The output of the 36-speed detector is conducted through L2 (fig. 16-4, A) and serves as the control winding current for the 36-speed magnetic amplifier. The control winding is loaded by a stabilizing inductance, L3, which establishes a phase-lag network. This alters the base of the signal to compensate for the effects introduced by the leading (capacitive) current resulting from the action of C2 and C4, respectively.

The one-speed phase detector has its input (reference and error voltages) paralleled with a similar phase detector, T4, used to supply direct current to the bearing (azimuth) error indicating meter on the front panel of the coder-indicator (chapter 15, figure 15-7). When the polarity of the input error voltage to T4 is as shown (fig. 16-4, A) voltages db and bc cause a current through CR13, R15, R16 and R17 in parallel, and CR16 to return to d. At the same time, voltages db and ba combine in series addition to cause a current through the movement of the azimuth error indicator meter (see figure 15-7). The current path is completed through the parallel combination of R16 and R17, which is in series with CR16. This current is proportional to the error signal at the indicator circuit input.

If the error voltage reverses its polarity with respect to the reference voltage at T4, voltages ab and bc will combine in series opposition to cause the current through the error indicator meter to decrease below the magnitude that would exist for zero error voltage.

ANTISTICKOFF VOLTAGE

As stated earlier, signals from the magnetic variation subassembly of the coder-indicator are introduced into the 1- and 36-speed differential transmitter in order to cause the pulser and plate assemblies in the antenna pedestal to be corrected to magnetic north. A single dial, (chapter 14, figure 17) calibrated from 0° to 360° and fastened to the coarse (1-speed) differential transmitter and control transformer (CDX1 and CT1 of figure 16-3) yields only a fair degree of accuracy in providing information to the servo motor, B1. A second dial, fastened to the fine (36-speed) differential transmitter and control transformer (CT2 of figure 16-3) is calibrated so that one revolution of the fine control transformer (CT2) equals only 10° of the full coverage of the coarse control transformer, CT1. Thus, a means of reading the position of the control shaft with 36 times the accuracy of a system with only the coarse control is provided.

The use of an even gear ratio between the fine and coarse control transformers presents a problem in synchronizing the servo system. The fine control transformer voltage (fig. 16-5, A) will be combined with the coarse control transformer voltage so that both voltages cross the horizontal axis simultaneously at 0° and 180°. At 0°, the output voltage of both the fine, and coarse control transformers is of the proper polarity to drive the servo to null. The 0 cross-over point is therefore said to be stable, since the two voltages cross the time-degree axis at the same time and in a direction with increasing amplitude. Note that at the 180° position on the time axis, the two voltages cross in opposite directions. Thus at 181° the 1-speed transformer voltage attempts to drive the servo motor, CCW through 179° to null at 360° or 0°, while the 36-speed transformer is driving the servo, CW, to null at 180°.

The fine control transformer voltage (from CT2 of figure 16-3) is fed to the servo only when the amplitude of the coarse control transformer voltage (from CT1 of this figure) is below 2.5 volts. Assuming that the servo is resting at a position 180° from its null position, if the 36-speed transformer voltage is then fed to the servo amplifier in the absence of a large one-speed transformer voltage, the servo would be synchronized at a false null of 180°. This difficulty is overcome by adding an a-c voltage, called antistickoff voltage, to the one-speed transformer voltage as shown in figure 16-5, B. The addition of this voltage prevents both error voltages from crossing the zero axis at the 180° point.

The antistickoff voltage is added to the one-speed voltage in the proper amount and with the proper phase to cause the one-speed voltage to be shifted 2.5° to the left of the zero point, and 2.5° to the right of the 180° point on the shaft angle axis. This corresponds to a 90° shift (or 1/4 revolution) of the 36-speed control shaft.

To correct the one-speed zero at the zero shaft angle position, the one-speed control transformer is re-zeroed to a point 2.5° from zero.

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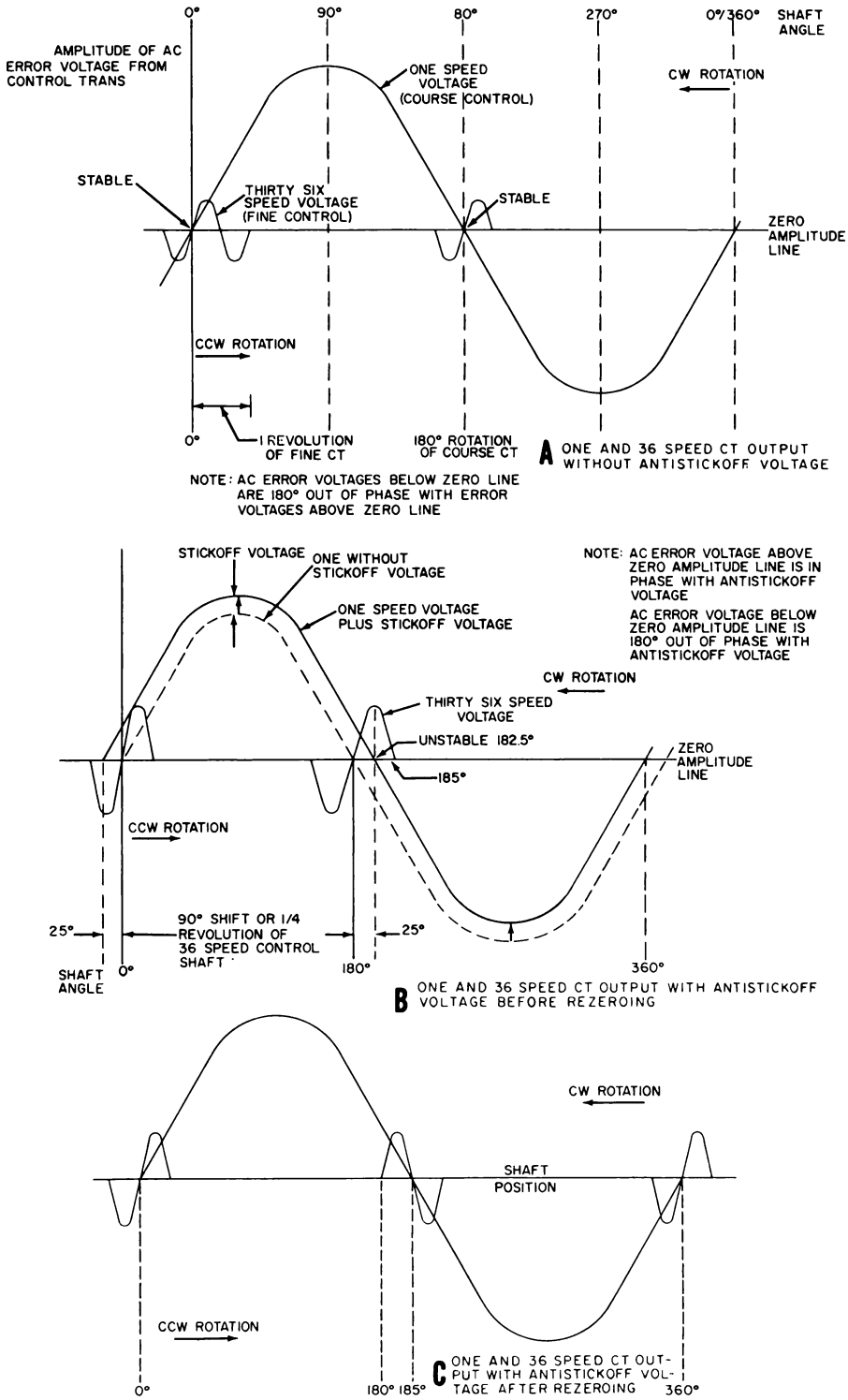

Figure 16-5.—A. Antistickoff voltage, one-speed and 36-speed control transformer output voltages, no antistickoff voltage added; B. One-speed and 36-speed synchro output voltages, antistickoff voltage added; C. One-speed and 36-speed output voltages with antistickoff voltage and one-speed synchro rezeroed.

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This action cancels the error introduced by the addition of the antistickoff voltage at the zero shaft position, and adds 2.5° error to the 180° crossover point.

Instead of passing through zero error voltage at the 180° point, the one-speed error voltage now passes through zero at 185° (180° plus 2.5° due to the antistickoff voltage, plus 2.5° due to re-zeroing of the one-speed synchro). Once this shifting takes place, the two voltages plus the antistickoff voltage appear shifted, as shown in figure 16-5, C.

Note that the antistickoff voltage is in phase with the one-speed error voltage from 0° to 180° of the shaft angle, therefore adding to the one-speed error voltage, and 180° out of phase with the one-speed error voltage from 180° to 360° of shaft angle, therefore subtracting from the error voltage.

The crossover point at 0° is now stable, while the other position of correspondence occurs at 185° and not at 180°. Therefore, the addition of the antistickoff voltage to the one-speed control transformer voltage eliminates the problem of false nulls in the servo system.

BEARING MAGNETIC AMPLIFIER (FEDERAL TELEPHONE AND RADIO CO. DESIGN)

As stated in chapter 14, there are two designs of the bearing magnetic amplifiers: the Magnetic Amplifiers Inc. Design (just discussed) and the Federal Telephone and Radio Company Design. The latter amplifier (fig. 16-6) provides power for the control phase of the bearing motor by means of a two-stage, push-pull magnetic amplifier, which is supplied from a phase sensitive detector.

A relay, K2, operated by a separate two-stage magnetic amplifier switches the input of the phase-sensitive detector from the 1-speed input to the 36-speed input when the 1-speed signal drops below a value corresponding to approximately 2.5°. The 36-speed output will result in a more accurate positioning of the pulser and plate assemblies.

The error signals at the stator of the 1- and 36-speed control transformers, CT1 and CT2, are fed to the amplifier circuit as shown in figure 16-6. With K1 energized from the 120-volt, 60-cycle supply, the one-speed error voltage is applied from terminal R1 of the one-speed control transformer (with the input polarity as shown) through the 8-9 winding of T1, R1, CR4, and the 5-4 winding of L10, through L1, and CR1, and through the closed contacts of K1, to return to R2 of the one-speed control transformer rotor. The phase of the error voltage (at the rotor) relative to the reference voltage (at terminals 8-9 of T1) at this time determines the amount of current through the 5-4 (control) winding of L10. These two voltages are in phase and additive for the polarities shown. This current is filtered by a 50-henry choke, L1, in series with the 5-4 winding. The mmf developed by the control (5-4) winding serves as the input to the switching preamplifier, L10.

Bias for the switching preamplifier is developed by the 6-7 winding of L10. That portion of the T2 secondary voltage (originated from the 120-volt, 60-cycle, single-phase input) induced into the 8-9 winding is bridge rectified by CR23, CR24, CR25, and CR26. Assuming the polarity shown, the current path is from terminal 8 of T2, through CR26, the series limiting resistors R2 and R3, the 6-7 winding of L10, and through CR24 of the bridge, to return to terminal 9 of T2. When the input cycle changes, the conduction path is the same except that the current passes through CR25 and CR23 of the bridge. Thus, a d-c bias current is applied to the 6-7 winding of L10.

The switching preamplifier, L10, and switching amplifier L11 are self-saturating reactor circuits, each employing two separate toroidal cores to control the energized and deenergized condition of a switching relay, K2. Before considering the effects of the control winding and bias current on the switching preamplifier circuit, study the quiescent (no error) circuit operation. Reference voltage is applied to the circuit from the 10-11 winding of T2. With the polarities shown for windings A and B of L10, and the 10-11 winding of T2, a current flows through the A winding of L10, through CR5, limiting resistors R4 and R5, filter choke, L2, the 4-5 winding of L11, and through CR7, to return to terminal 11 of T2. During the next half-cycle, winding B of L10 will conduct and current flows through the 4-5 winding of L11 in the same direction.

During normal operation, the 4-5 control winding of L10 (in the presence of the bias developed by the 6-7 winding, as discussed) controls the magnitude of the voltage developed at A and B during the conduction period. Thus, the

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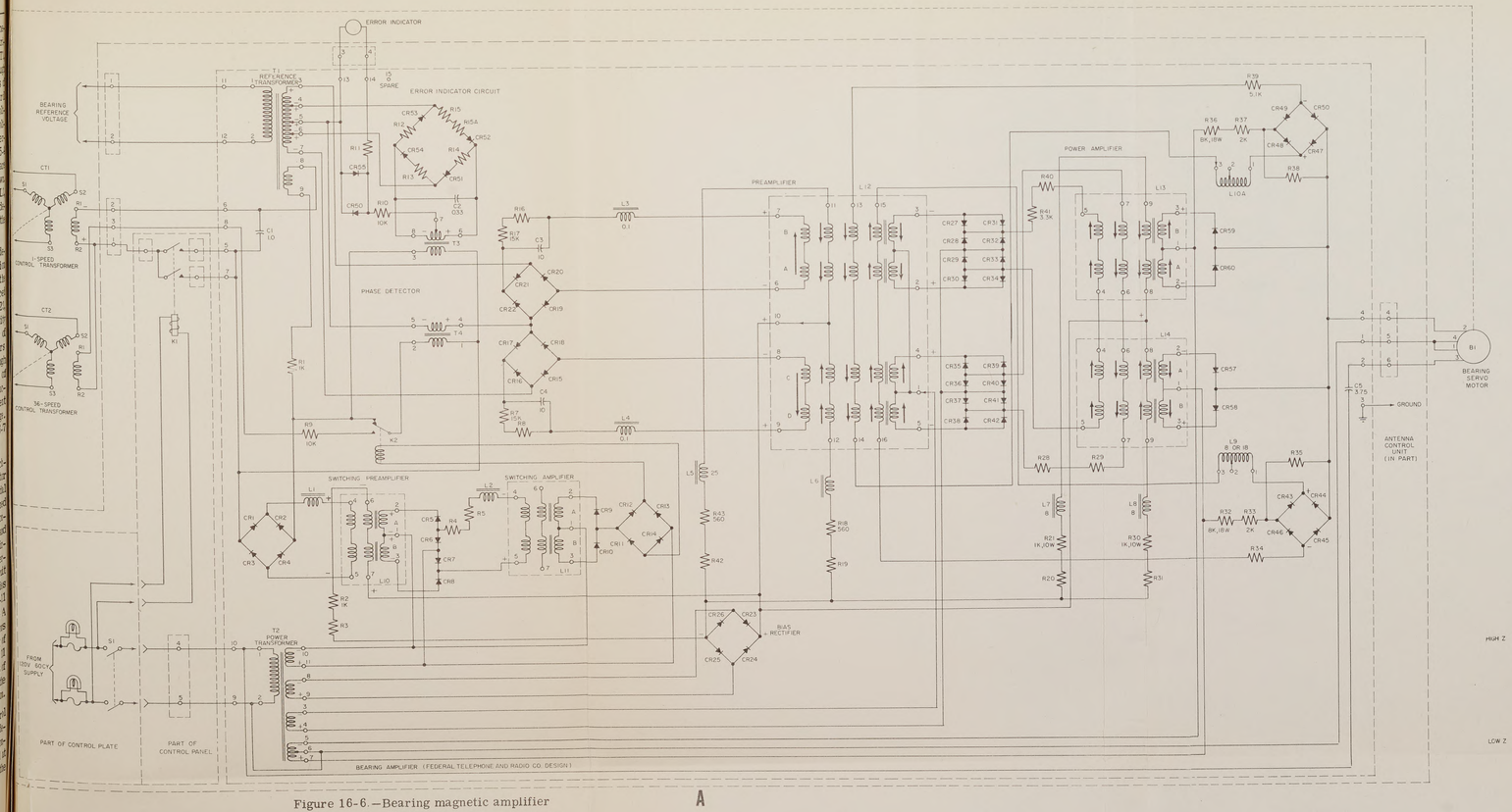

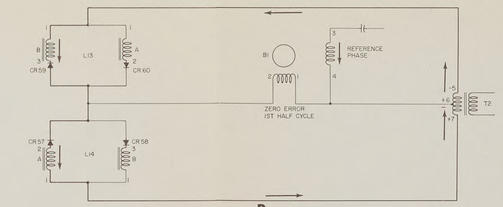

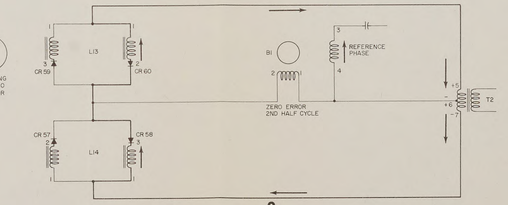

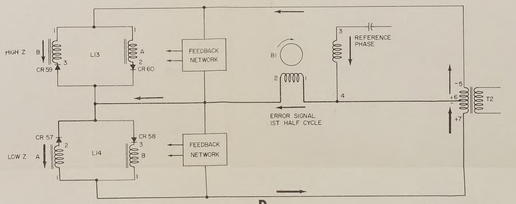

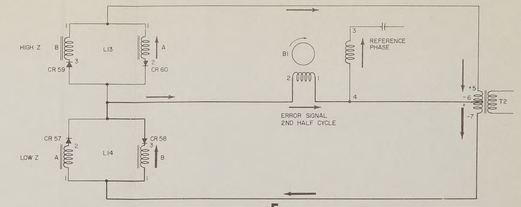

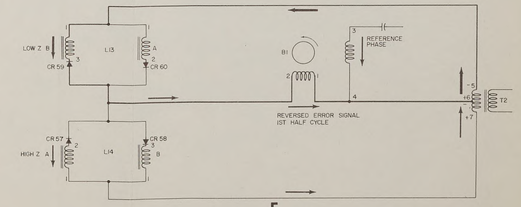

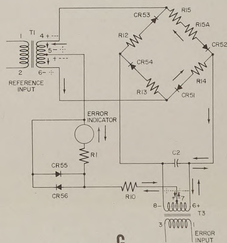

Figure 16-6.—Bearing magnetic amplifier (Federal Telephone and Radio Co.). Schematic diagram with equivalent bridge output circuits and error indicator circuit.

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itput current in the 4-5 winding of L11 is aused to vary in magnitude.

The switching amplifier, L11, under the inuence of the 4-5 control winding and the referace voltage from the 10-11 winding of T2, delops an output, which is rectified by the bridge ircuit comprising CR11, CR12, CR13, and CR14. 11 does not receive a bias current through the -7 winding.

The bias mmf (6-7 winding) of L10 is in opposition to the CT1 output mmf (4-5 winding) so iat the degree of saturation of L10 depends upon ie difference between the two mmf's. With zero rror signal from CT1 the 4-5 winding mmf is mall, and the bias (6-7 winding) of L10 satuates the core. Thus, with little flux change the mpedance of the 2-3 load winding is small, and ie output voltage to the 4-5 control winding of 11 is relatively large. Large current through iis winding results in core saturation of L11 ith relatively low impedance developed in the 11 2-3 load winding. Thus, the rectified voltage nd current of relay K2 are relatively large, and 2 closes its lower contact.

In the presence of an error signal less than .5° to the 4-5 control winding of the switching reamplifier, L10, the controlling current to the witching amplifier L11 permits core saturation f L11 and again K2 will operate to close its ower contact. For large inputs (resulting from rror signals 2.5° or more) the L10 bias is eutralized, and the L10 core does not saturate. hus, the impedance of the L10 2-3 winding is elatively large. The resulting current through ie L11 4-5 control winding is small, and the L11 ore is not saturated.

The impedance of the L11 2-3 load winding s relatively large, and its current is not suffiient to hold K2 in the energized position. Thus, 2 opens its lower contact and closes its upper ontact.

Since the input path to the 36-speed circuit rom CT2 to T4 in the phase detector will be pened at the lower K2 relay contacts (in the eenergized position), only the one-speed output rom CT1 to T4 in the phase detector is effective controlling the servo motor for error signals reater than 2.5°.

When the error signal drops below 2.5°, the iput voltage is small, and K2 is energized, as reviously described. The 36-speed error input s applied from terminal R1 of CT2 through R9 nd the lower contacts of K2, the primary of T4, ie closed contacts of K1, to return to terminal 2 of the 36-speed control transformer. Thus, the 36-speed output alone is fed through the servo amplifiers to the servo motor for error signals less than 2.5°.

The phase detector circuit comprises two bridge rectifiers, which supply currents through equal portions of the control windings 6-7 and 8-9 of the preamplifier, L12, circuit in opposite directions. The circuit sums up the phase of the error input (at T4) and the reference voltage input (at T1) to yield a controlling current through the output winding, which is proportional to the phase of the two signals.

Reference voltage for the preamplifier circuit is obtained from the 3-5-7 terminals of T1.

Assuming that there is no error signal input and a reference voltage input of the polarity shown at the T1 secondary, a quiescent current will flow through from terminal 7 of T1, through CR15, the 8-9 winding of L12, L4, R8 and R7, CR17, CR19, the 6-7 winding of L12, L3, R16, and R17, CR21, returning to terminal 3 of T1. This value of current represents the quiescent control winding current to the preamplifier, L12.

When an error signal is introduced, the phase of the signal will cause the error input to be additive to one-half of the T1 secondary voltage and subtractive to the other half. If the polarity of the error voltage is as shown, the voltage at the 5-7 winding of T1 will add to the error voltage of the T4 5-4 winding to produce a relatively large current through CR15, the 8-9 winding of L12, through the filter choke, L4, through limiting resistors R7 and R8 (bypassed by a filter capacitor, C4) and through CR17 and the T4 secondary to complete the circuit. The 3-5 winding voltage of T1 is in series opposition to the 4-5 winding voltage of T4 with the indicated polarities. The path for the 6-7 winding current for this half cycle is from terminal 5 of T1, the 5-4 winding of T4, CR19, the 6-7 winding of L12, L3, R16, R17, CR21, and terminal 3 of T1. The 6-7 winding current of L12 is smaller than the 8-9 winding current of L12 for the indicated polarities.

Conversely, if the polarity of the input error voltage at this time is opposite to that shown, the input voltage would be additive to the potential present at the 3-5 winding of T1, and the current through the top bridge (and the 7-6 winding of L12) would be the larger of the two currents. This current (and the corresponding mmf) is in opposition to that in the 8-9 winding of L12. The action of the phase detector circuit is the same for

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both the 1- and 36-speed inputs, which, as discussed, is dependent upon the postion of the K2 contacts.

Bias for the input of the preamplifier, L12, is obtained from the bias bridge rectifier comprising CR23, CR24, CR25, and CR26. The operation of this circuit was considered in the discussion of the switching preamplifier. The bias for the two sections of the bias winding is applied to the 11-10 and 12-10 windings in parallel from the CR25-CR26 and CR23-CR24 junctions. In this way the bias for the 10-11 winding is in opposition to the bias in the 12-11 winding. R42 and R43 are current limiting resistors for the top bias circuit, while R18 and R19 serve the same purpose for the bottom circuit. L5 and L6 are filter chokes for their respective circuits.

Windings 13-14 and 15-16 are feedback windings. The operation of these windings to control the preamplifier circuit is described later.

As previously stated, the 7-6 and 8-9 control windings of L12 produce mmf's, which are always in phase opposition. Further, it has been established that when the current through the 7-6 winding increases there is a corresponding decrease in the current through the 8-9 winding of L12. The portion of the control circuit that conducts the heavier current is determined by the phase of the error signal input.

The preamplifier, L12, is divided into two equal sections, which are push-pull connected. If we assume a quiescent (no error input) condition with the opposite polarity of the voltage between the 3-4 terminals of the T2 secondary (shown in dotted lines), a current will flow in the 1-5 and 1-3 windings of the preamplifier, L12. The conduction path is from terminal 4 of T2, through CR36 and CR40 (in parallel), the 7-6 windings of L13 and L14, R29 and R28, CR38 and CR42 (in parallel), and the 5-1 winding of L12, to return to terminal 3 of T2. The top section of the preamplifier conducts from terminal 4 of T2, through CR29 and CR33 (in parallel), the 5-4 windings of L13 and L14, R40 and R41, CR27 and CR31 (in parallel), and the 3-1 winding of L12, returning to terminal 3 of T2. Thus, for this half cycle during quiescence, a current flows through the separate control windings, 4-5 and 6-7 of L13 and L14, which are of equal magnitudes but opposite directions.

During the next half cycle of the input (polarity shown in solid lines) the conduction path is the same except that the L12, 1-4, and 1-2 windings replace the 1-5 and 1-3 windings, and the conducting diodes include CR35, CR39, CR37, and CR41 for the bottom section of the preamplifier, and CR30, CR34, CR28 and CR32 for the top section of the preamplifier. The diodes are parallel connected to increase the current-carrying capabilities of the circuit.

The current through the control windings, 4-5 and 6-7, of L13 and L14, as discussed, are in phase opposition (opposing mmf's) and represents the controlling mmf's for the power amplifiers. Since the two mmf's are equal and opposite, the resultant mmf is zero when the error input signal is zero.

The bias mmf in the 11-10 winding of L12 is always in opposition to the control mmf in the 7-6 winding of L12. Likewise, the bias current in the 12-10 winding of L12 is always in opposition to the control mmf in the 8-9 control winding.

The bias mmf in the 11-10 winding of L12 is in the same direction as the load mmf in the 3-2 load winding of L12. Thus, with no error signal input from T3 at the 7-6 control winding (mmf weak), the bias mmf of the 11-10 winding tends to saturate the L12 core, and the impedance of the 3-2 winding is low. Likewise, the bias current in the 12-10 winding of L12 produces an mmf, which aids the mmf of the 5-4 winding of L12. Thus, the impedance of the 5-4 winding is also reduced because of core saturation in the presence of a weak 8-9 control winding mmf.

Assume that an error signal is introduced at T4, which results in a larger current through the B-A winding of L12 and a smaller current through the C-D winding. The larger mmf produced in the 7-6 control winding has a greater opposing influence on the bias (11-10) winding, and the impedance of the 3-2 load winding is greatly increased. Similarly, the impedance of the 4-5 load winding of L12 will also increase but not as much. Since the impedance of the top half of the preamplifier is increased a large amount, a smaller magnetomotive force will be developed in the 5-4 winding of L13 and L14 than in the 7-6 windings.

A bias mmf is established in the 9-8 windings of L13 and L14 (power amplifier) from the bias rectifier in the same manner as described for the preamplifier, L12. Note that the bias mmf in the 9-8 windings of L13 aids the mmf in the 7-6 control windings of L13, but opposes the mmf in the 5-4 control winding. Conversely, the bias in the 9-8 winding of L14 opposes the mmf in the 6-7 control winding, but aids the mmf in the 5-4 control winding of L14.

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Before considering the effects of an error signal introduced at T4, note the action that takes place in L13 and L14 for zero error. For zero error, the currents are equal in the two control windings (7-6 and 5-4 windings of L13 and L14), and the two control mmf's cancel in the four cores of the power amplifier. Also, with zero error, the bias mmf's in the 9-8 windings of L13 and L14 oppose the 3-2-winding mmf's by equal amounts, and both 3-2 load windings of L13 and L4 have equal values of high impedances and restrict the output currents to small equal values.

Consider the action of the 3-2 load windings of L13 and L14 in the simplified equivalent bridge circuit (fig. 16-6, B). When the error signal is zero, the impedances of the 3-2 load windings of L13 and L14 are equal, and the bridge is balanced. The control phase winding, 2-1, of motor B1 receives no current because the voltage across this portion of the bridge is zero.

On one half cycle current flows from terminal 5 of T2 with winding 1-3 of L13 and winding 2-1 of L-14 conducting with the current returning to terminal 7 of T2. On the second half cycle (fig. 16-6, C) winding 1-3 of L14 and winding 2-1 of L13 conduct. The motor does not rotate because only the reference phase is energized.

Now consider the effect of an error signal from T4 on the power amplifier output. It was previously assumed that the current in the A-B control winding of the preamplifier, L12, was greater than the current in the C-D preamplifier winding. From this it was determined that the impedance of the 3-2 load winding of L12 became much greater than the impedance in the 4-5 load winding of L12. Thus, the mmf produced in the 5-4 control windings of L13 and L14 is much less than the mmf developed by the 7-6 windings of L13 and L14 .

Under this condition, a predominant mmf is generated in the 7-6 control winding of L13, which opposes the mmf in the 3-2 load winding of L13. The impedance of the 3-2 load winding of L13 is therefore increased. The 7-6 control winding mmf of L14 is also predominating, but aids the 3-2 load winding mmf of L14. This action reduces the impedance of the 3-2 load winding of L14 because both cores A and B tend to saturate. Thus, the over-all effect of the assumed error is to increase the impedance of the 3-2 load winding of L13, and to decrease the impedance of the 3-2 load winding of L14.

The bridge circuit (fig. 16-6, D) is now unbalanced because the L14 impedance is low, and the L13 impedance is high. The current through the 6-7 winding of T2 now exceeds the current in the 5-6 winding of T2. For the first half cycle the current in the lower lead (heavy arrow) flows into terminal 7 of T2 and divides at terminal 6 of T2. A portion of this current flows out from terminal 6 of T2 and through the control winding 1-2, of the motor, B1, to produce motor torque. The direction of the current in the top lead is the same as the middle lead of the bridge for this half cycle (winding 1-3 of L13 and winding 2-1 of L14 conducting). On the second half cycle (fig. 16-6, E), the polarities reverse in the control and the reference phases, but the direction of rotation of motor B1 remains the same. On this half cycle the 1-3 winding of L14 and the 2-1 winding of L13 are conducting.

If the error signal from T4 reverses its polarity with respect to the reference phase polarity, the L13 load winding impedance will become small, and the L14 load winding impedance will become large (fig. 16-6, F). For the first half cycle the direction of the unbalanced current through the 2-1 control winding of motor B1 is reversed with respect to its direction in figure 16-6, D, and the reference phase so that the direction of rotation of the motor is reversed.

Two feedback networks (represented by the blocks in figure 16-6, D) are connected across each half of the power amplifier output circuit. These networks rectify and filter a portion of the total power amplifier output. The resulting current from each network is used as a negative feedback to the preamplifier, L12, to prevent overdriving the servo motor.

The top feedback circuit (fig. 16-6, A) comprises limiting resistors R39, R38, R36, and R37 and bridge rectifiers CR47, CR48, CR49, and CR50.

This bridge conducts a current (from the negative bridge terminal) downward through the 13-14 feedback winding of L12. A similar bridge in the bottom section of the power amplifier conducts a current upward through the 16-15 winding of L12. Thus, the two feedback windings (13-14 and 15-16 of L12) develop opposing mmf's in the four cores of the preamplifier.

When the impedance of the top power amplifier, L13, is increased (as will be the case for the error assumed at the A-B and C-D windings of preamplifier L12) the bridge of figure 16-6, D, is unbalanced (as discussed), and the voltage across the top feedback network will be larger than the voltage across the bottom feedback circuit because the impedance of L13 is higher

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than that of L14. Thus, the current conducted downward through the 13-14 feedback winding of L12 (fig. 16-6, A) predominates. This action tends to decrease the impedance of the 3-2 load winding of L12 while simultaneously increasing the impedance of the 4-5 load winding of L12. Since this action is opposite to the effect caused by the B-A and C-D control windings of L12, a degenerative (negative) feedback is produced, which prevents overdriving the servomotor in the direction dictated by the error input.

L9 and L10 are filter chokes for the feedback currents. Each of the chokes may be tapped at 8- or 18-henry terminals. The power amplifier output is applied between terminals 2 and 1 of the motor, B1. Reference phase voltage is applied from T2 (when S1 is in the closed position) to terminals 4 and 3 of the motor. C5 is a phase shift capacitor for this voltage.

An error indicator circuit receives the one-speed input via T3 (fig. 16-6, A). The input connection is made ahead of the switching relay so that the one-speed input is available at the error indicator at all times.

The circuit of the error indicator is that of a phase sensitive detector. Reference voltage for the error indicator circuit is obtained from the 4-6 terminals of T1. This potential is applied across a bridge circuit comprising di[illegible]R51, CR52, CR53, and CR54 and limiting [illegible]s R12, R13, R14, R15, and R15A. The latter two resistors are selected by the manufacturer for alignment purposes and may be omitted in some cases.

The circuit of the error indicator is explained with the use of figure 16-6, G. In the absence of an error input (with the instantaneous polarities at T1 and T3, as shown in solid lines), a current is conducted by T1 from terminal 6 of T1, through CR51, R14, CR52, R15A, and R15 to terminal 4 of T1, while a current is also being conducted by T3 from terminal 8 of T3, through CR54, R13, CR51, and R14 to terminal 6 of T3. During the next half cycle, the current is reversed through the bridge.

For each half cycle of the input, a current will also be conducted through the error indicator meter movement. For the solid line polarities at T1 and T3, a current is conducted through the meter from terminal 7 of T3, through R10, R1, the error indicator, the 5-6 terminals of T1, CR51, and R14, to return to terminal 6 of T3. During the next half cycle, all polarities at T1 and T3 are reversed, and electrons flow from terminal 7 of T3, through R10, R1, the error indicator, the 5-4 terminals of T1, CR53, and R12, to return to terminal 8 of T3. Note that the current through the meter is in the same direction for both half cycles of the input, and the azimuth error indicator meter (chapter 15, figure 15-7) reads at a central zero error position.

If an error signal is introduced at T3, which aids the voltage shown in solid lines, the current through the meter will be increased, and the meter will read up-scale from its center position. Conversely, if the introduced error opposes the voltage at T3 (indicated in dotted lines), the meter current will be decreased, and the meter will read downscale from its zero center position. CR55 and CR56 are placed across the meter movement and conduct when the error voltage exceeds a value that may be harmful to the meter.

ROLL AND PITCH SERVOSYSTEMS (MAGNETIC AMPLIFIERS, INC.)

Roll and pitch stabilization control facilities are employed in the shipboard radio beacon to compensate for errors introduced into the beacon information by movements of the ship. These facilities are not used with shore-based antenna groups because stabilization of the fixed antenna is not required.

Roll and pitch information is fed from the ship's bus to the antenna stabilization servo system (fig. 16-3) through the select antenna position switch, S1, and their respective control transformers, CT3 and CT4. S1, which is located on the panel of the coder-indicator (figure 7, chapter 15) may be placed in either a STABILIZED or STOW position. For shore installations, S1 is placed in the STOW position, and for shipboard installations, the switch is placed in the STABILIZED position.

The error signal developed by the separate control transformers is fed to its respective magnetic servo amplifier (either roll or pitch). The magnetic amplifier, in turn, supplies power to the servo motor (B2 or B3 of figure 16-3), commanding a specific direction of rotation determined by the error signal.

As stated in chapter 14, the servo motor, through its associated gear train, moves the antenna to its corrected position. At the same time, the gear train nulls the error signal at the control transformer through a 1:2 gear box train (fig. 16-3).

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A rate feedback loop is provided through respective roll or pitch tachometers (TC1 and TC2) to eliminate oscillations. A tachometer is an instrument for measuring angular speed in revolutions per minute. The tachometer feedback improves the overall stability of the servo system. A detailed consideration of rate feedback will be presented later.

The magnetic amplifiers for the roll and pitch are identical electrically. The operation of these circuits is similar to that of the bearing magnetic amplifier already discussed.

Roll and pitch magnetic amplifiers of two different designs may be furnished with the equipment. One type is manufactured by Magnetic Amplifiers, Inc., and the other by the Federal Telephone and Radio Company. Both designs are treated in this discussion.

The roll amplifier and pitch amplifier are almost identical units. Accordingly, this discussion applies to both units, except where explained otherwise. Each amplifier (either roll or pitch, figure 16-7) is composed of a subassembly, which contains a phase-sensitive detector (not shown), a magnetic amplifier, a saturable reactor (fig. 16-8), a power transformer, and two motor capacitors (one shown).

The synchro error voltage, which is obtained from the roll and pitch control transformers (located in the antenna base) is fed into a phase-sensitive detector. The phase detector is not shown, but is of the type considered in the discussion of the bearing magnetic amplifier. The d-c output serves as the control current for the preamplifier (fig. 16-7).

The control current from the phase-detector passes through three pairs of control windings, which are connected in series aiding. In addition, a tachometer (fig. 16-3) supplies a feedback current (from TC1 for pitch and TC2 for roll) through two of the control windings in series. Isolation of the two input circuits is obtained by the network impedance of C2, R3, and R4 in series with C1 in the tachometer input circuit, and gain control resistors, R1 and R2 in the control current circuit.

The gain of the roll and pitch amplifier (which is controlled by R1 and R2 in parallel) is preset at the factory, as required for each amplifier. The isolation impedance in the tachometer circuit serves the purpose of changing the tachometer voltage from rate feedback to acceleration feedback. Rate and acceleration feedback are discussed in the following paragraphs.

When the load is turning at a constant speed, a steady d-c voltage (proportional to speed) is fed back to the preamplifier in opposition to the error signal. Therefore a greater error signal is required to overcome the d-c voltage and produce a certain unbalance in the amplifier and a certain speed of the servo motor than would be required if there were no feedback.

In order to produce a greater error signal there must be a greater error. Since the feedback voltage is proportional to speed, the error is also proportional to speed. This is called a velocity error, or rate feedback.

To eliminate this velocity error, a capacitor, C1, is connected in the feedback circuit. The capacitor blocks the d-c component of the voltage. The a-c component of the voltage at the frequency of hunting (several cycles per second) is fed back without much loss. (The capacitor has a high capacitance and low reactance). Now, only changes in speed will produce a feedback signal through the capacitor. This action is called acceleration feedback, which further improves the stability of the servo system.

Damping the servo system by feedback without the capacitor is analogous to damping a synchro by putting friction in the bearings; whereas, the feedback circuit with the capacitor is analogous to the synchro flywheel damper used in synchro receivers.

Bias current for the preamplifier is obtained from the 18-19 and 28-29 windings of L1. These windings are connected in series with separate bridge rectifier currents across the secondary of power transformer T1. The current conducted through the load is determined by the degree of saturation of the core material of L1. This current serves as the control and bias current for a magnetic amplifier.

Saturable reactor L2 (fig. 16-8), along with two feedback bridge rectifiers and a power transformer, comprises the output magnetic amplifier, which feeds the roll and pitch motors. T2 supplies power for the two bridge rectifiers. C2 is a power factor correction capacitor for the control phase of the motor. Another capacitor (not shown) is connected in series with the line, which supplies the reference phase of the motor to provide the phase splitting required for starting the motor.

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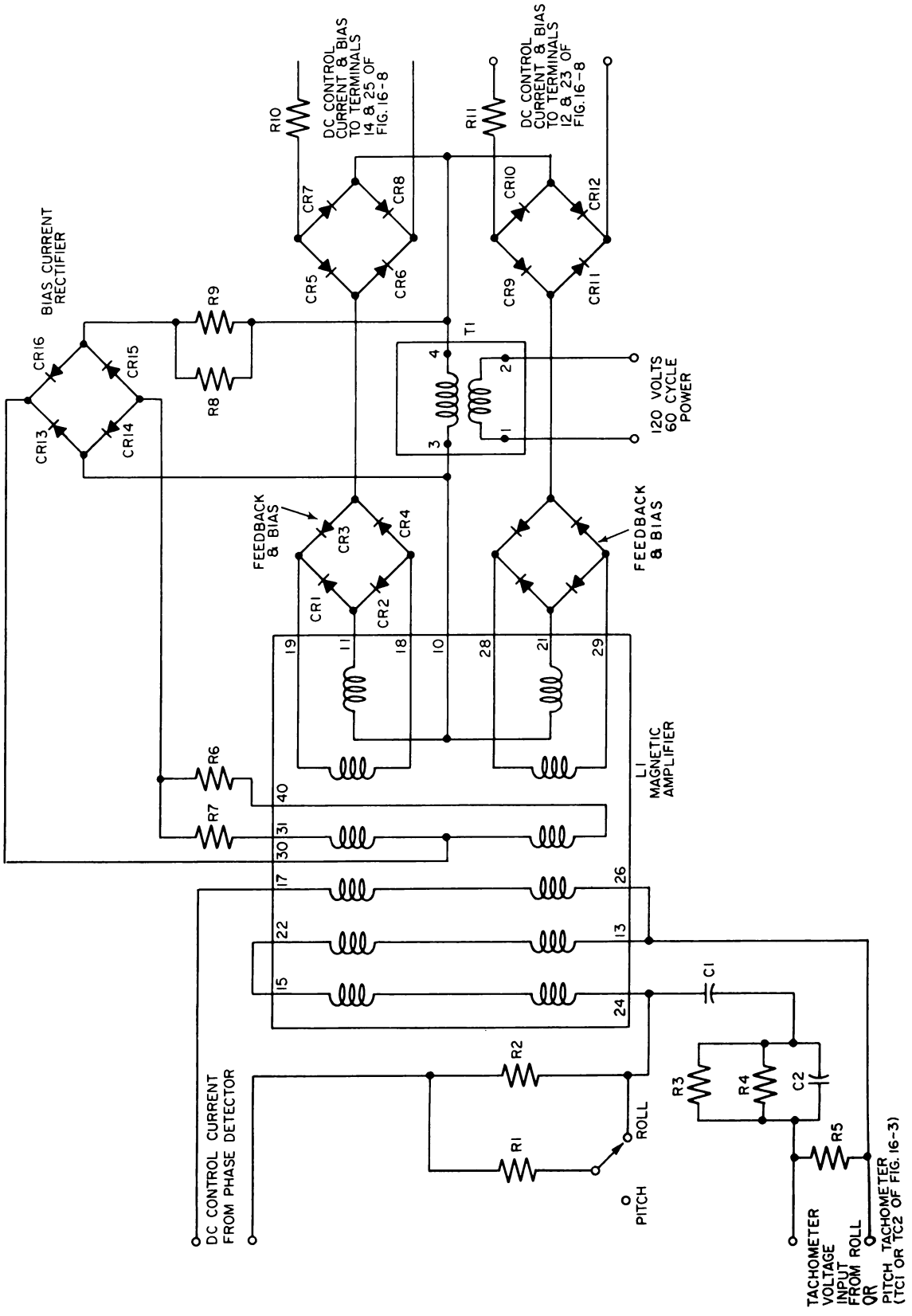

Figure 16-7.—Roll or pitch magnetic preamplifier (Magnetic Amplifiers, Inc. design).

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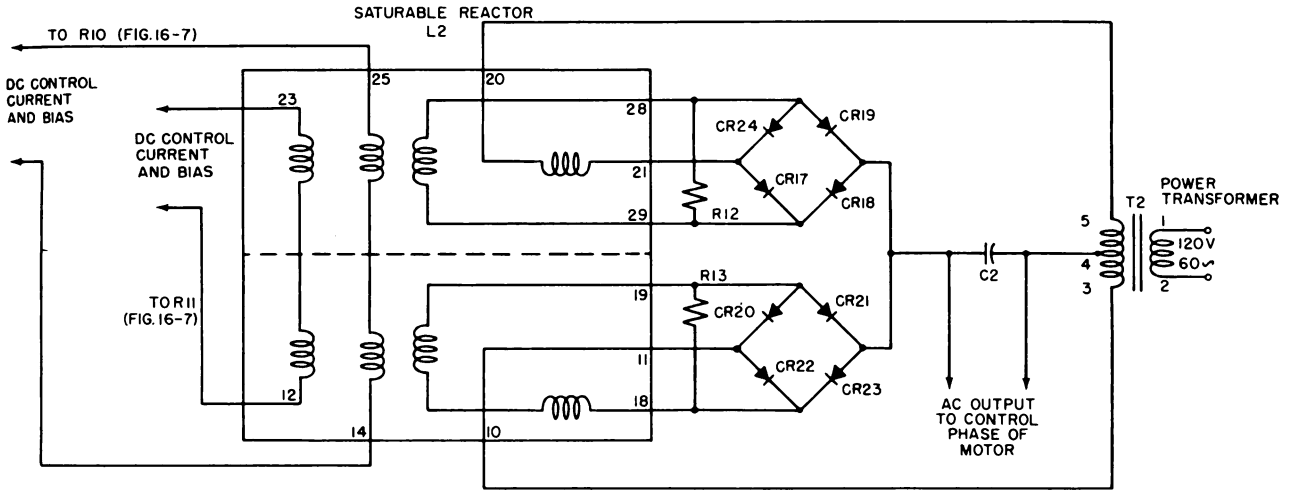

Figure 16-8.—Roll or pitch magnetic power amplifier (Magnetic Amplifiers, Inc. design).

ROLL AND PITCH SERVO SYSTEMS (FEDERAL TELEPHONE AND RADIO CO.)

The roll amplifier and pitch amplifier of the Federal Telephone and Radio Company Design (fig. 16-9) are identical to each other. The amplifiers provide power for the control phase of the roll or pitch servo motor (B2 or B3 of figure 16-3), commanding a specific direction of rotation of the motor required to correct the antenna position.

Power for the roll or pitch servo motors is supplied by means of a three-stage arrangement of push-pull magnetic amplifiers (fig. 16-9), which are fed from a phase sensitive detector. The three stages of magnetic amplification are the preamplifier, the booster, and the power amplifier. Because these circuits are essentially the same as those described in the bearing magnetic amplifier (figures 16-4 and 16-6), only a brief discussion of the circuit operation will be presented. While following the discussion, keep in mind that the arrows in the illustration (fig. 16-9) indicate the direction of the magnetomotive forces.

The synchro error voltages from the roll or pitch control transformers (CT3 or CT4 of figure 16-3) are applied to the phase sensitive detector (fig. 16-9) via the secondary of T3. The reference voltage is applied across T1. If the incoming error signal is in phase with the reference voltage, the output of the rectifier comprising CR7 through CR10 will be increased. Since the phase detector output current flows in opposition to the output current in the 3-2 load winding of L1, the top preamplifier output will be decreased. Similarly, the output of the bottom preamplifier will be increased.

Both top and bottom preamplifier outputs are applied to the booster amplifier comprising L2 and L3. As the output of the top preamplifier (3-2 load winding of L1) decreases, the output of the top booster amplifier, L2, will increase. Similarly, the output of the booster amplifier, L3, will decrease.

The output of the booster, L2, is applied to its corresponding power amplifier, L4, while the output of booster L3 is applied to L5. In the power amplifier (L4 and L5) the mmf in the principal control winding (6-5 windings of L4 and L5) aids the mmf in the 3-2 load winding of L4 and L5. Thus, an increased output of either section of the booster amplifier causes an increased output of the corresponding power amplifier.

For zero error, L4 and L5 produce equal outputs. If the booster outputs are such that L4 and L5 are thrown out of balance, the phase of the larger output predominates, and is fed from the center-tap of the 3-2 load windings of L4 and L5 to the control phase of the servo motor (fig. 16-3).

A stabilizing feedback to the booster (fig. 16-9) is obtained by rectifying the a-c voltages, which appear across the power amplifier 3-2 load winding. The feedback current is fed through the 12-11 and 14-13 windings of L2 and L3, respectively.

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The tachometer feedback signal to the roll or pitch amplifiers is originated at the servo motor (B2 or B3 of figure 16-3) and fed through a 1:2 gear box to the tachometer (TC1 or TC2). The tachometer feedback signal is fed through the tachometer feedback network (fig. 16-9) comprising R1, R2, R3, R4, and C1, to the 10-11 winding of L1. C1 in the feedback network is effective in changing rate feedback to acceleration feedback, as discussed.

A second phase detector in the roll and pitch amplifiers (fig. 16-9) is used in the error indicator circuit. The circuit of the error indicator is identical to that described in the discussion of the bearing magnetic amplifier (fig. 16-6 G). The error indicator meter for the roll and pitch amplifiers is located on the front panel of the coder-indicator (chapter 15, figure 15-7).

SERVO MOTORS

The servo motors are 100-watt, two-phase motors that have a continuously excited phase and a variable phase. Connected to the roll and pitch servo motors (fig. 16-3) are blowers (not shown), which help dissipate the heat of the units. The bubble canopy (radome in chapter 14, figure 14-1) protects the antenna and base from the weather, and serves to decrease wind resistance. Without the bubble canopy, much larger servo motors would be required.

ANTENNA SPEED CONTROL

As discussed previously, the speed of the rotating cylinders carrying the parasitic reflectors must be maintained at a constant 900 rpm. This is true because the speed of the antenna controls the frequency of modulation of the beacon output.

The antenna speed is maintained constant through the use of a tachometer, TC3 (fig. 16-3) in a speed control circuit. The tachometer is geared to the speed control motor, B4, and generates 810 cps when the antenna cylinders are spinning at the required 900 rpm.

The speed control amplifier circuit (fig. 16-10) receives the tachometer output (from TC3 of figure 16-3) at a frequency sensitive circuit comprising T1, Z1 and Z2. This circuit is essentially a frequency discriminator which senses the error in the tachometer frequency from the normal 810 cps, and produces two output currents whose difference is proportional to the frequency error. The frequency sensitive characteristics of the circuit are obtained by means of two series resonant tanks (Z1 and Z2, respectively); whereas, Z1 is tuned to 805 cps and Z2 is tuned to 815 cps (see figure 16-11). T1 and T2 are coupling and impedance matching transformers.

The output of the two circuits is rectified by separate bridge rectifiers connected across their respective sections of the output transformer, T2. The variation of the direct current output of the bridge rectifier for the lower frequency resonant circuit as the tachometer frequency is changed, as illustrated in figure 16-11, A. The d-c output of the higher frequency resonant circuit for a changing tachometer input is illustrated in figure 16-11, B. Note that the two resonant curves are displaced equally on either side of the nominal tachometer frequency. A curve representing the difference of the two curves, which is the error curve that effectively controls the speed of the antenna spin motor (fig. 16-3) is shown in figure 16-11, C. The difference current is zero at 810 cps, and only when a deviation from this frequency occurs will there be an error current to the spin motor (B4 of figure 16-3), which has the correct sense to return the speed of the antenna cylinders to 900 rpm ± one-half percent.

Opposing mmf's are produced in the control windings of saturable reactor, L1 (fig. 16-10) by the high and low frequency rectifier circuits. Actually the difference current illustrated at C of figure 16-11 does not exist in the amplifier, but is effectively produced by the opposing mmf's at the 12-13 and 14-15 control windings of L1. R2 and R3, in conjunction with C1 and C2, form a phase lead network to compensate for the inductive (lagging) current in the lower frequency circuit caused by the reactance of the control winding. A similar network is connected in series with the control winding of the higher frequency circuit. L2is a stabilizing (loading) inductance, which is necessary for eliminating excessive overshoot of the antenna cylinder speed.

The T3 secondary voltage is rectified to supply a d-c bias current to the 1-2 winding of L1. The output of L1 is taken from the 10-11 winding and rectified by the bridge circuit composed of CR17, CR18, CR19, and CR20. A portion of this putput is rectified and used as positive feedback in the 18-19 winding.

When the error signal is zero, a rectified current from the T3 secondary is conducted via

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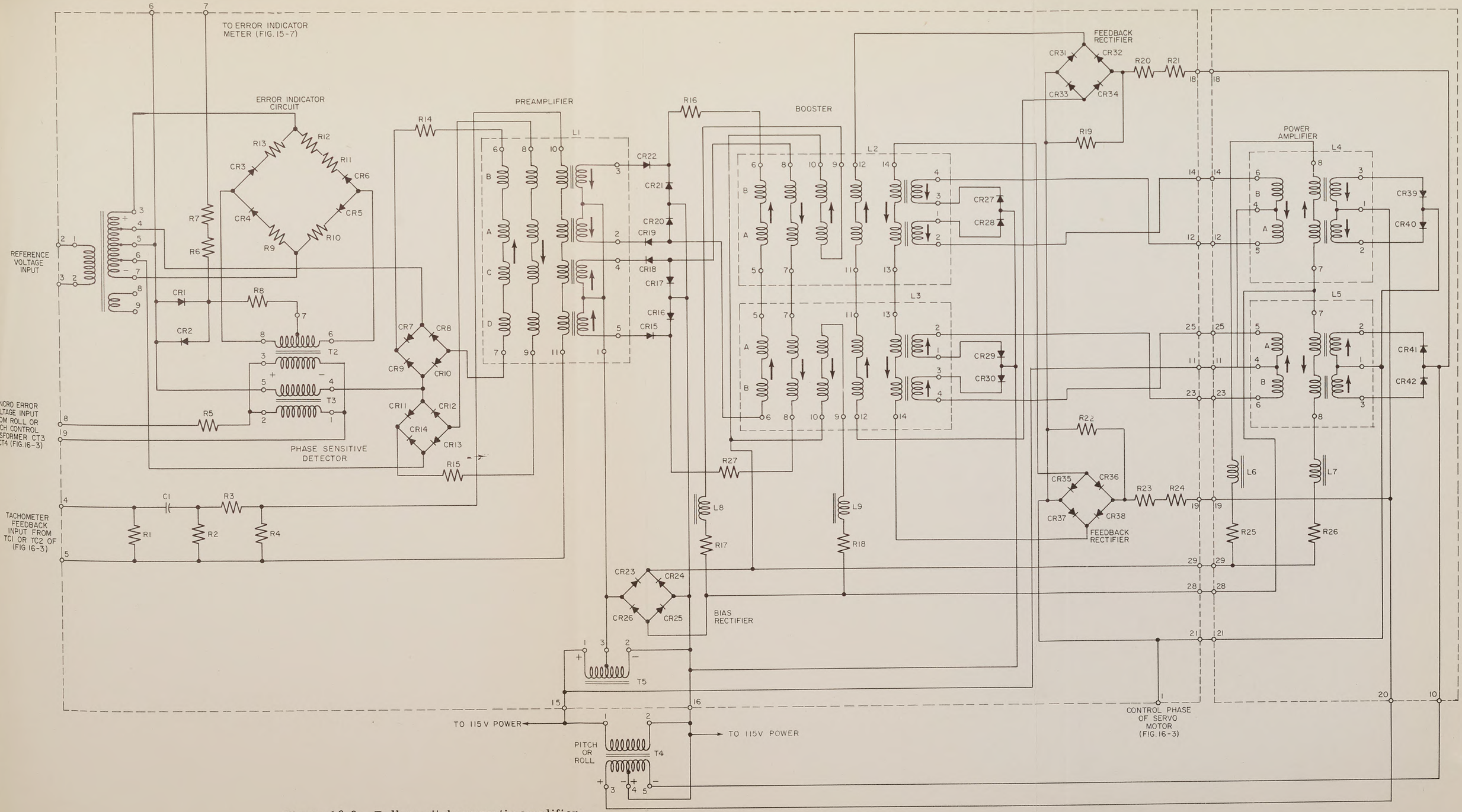

Figure 16-9.—Roll or pitch magnetic amplifier (Federal Telephone and Radio Co.).

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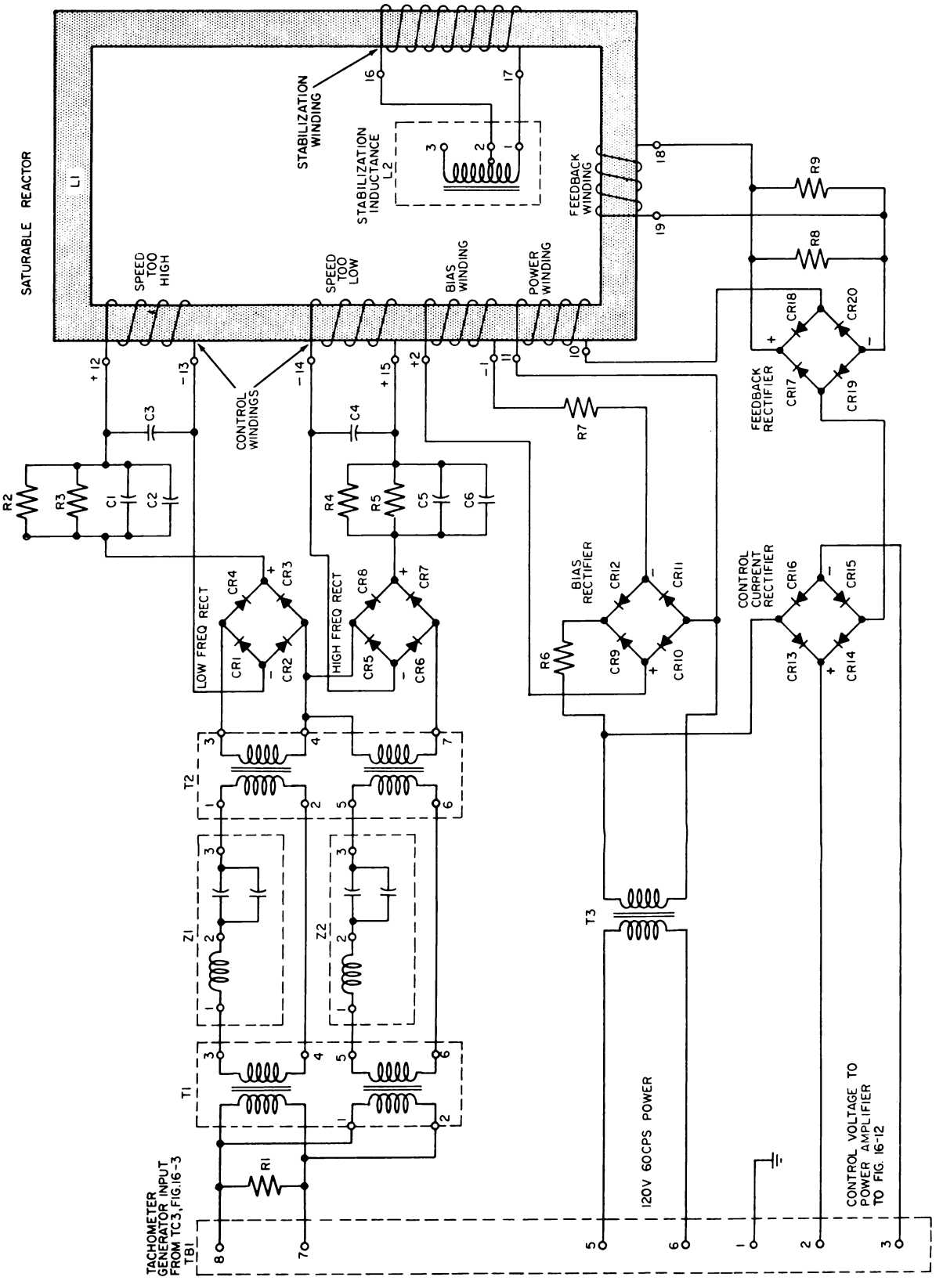

Figure 16-10.—Antenna speed control preamplifier circuit.

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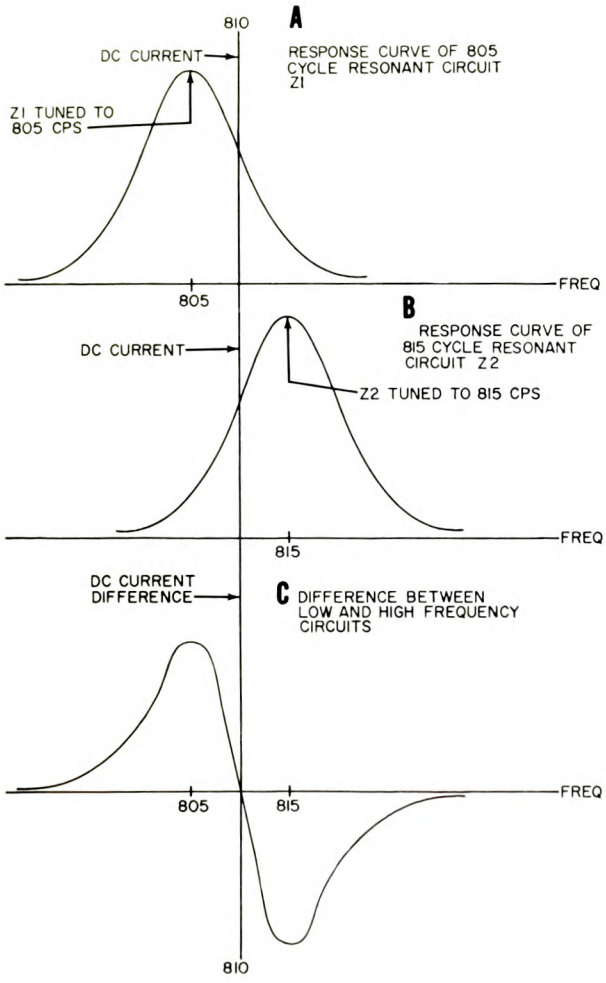

Figure 16-11.—Antenna speed control circuits, analysis of waveshapes.

terminals 2 and 3 of TB1 through the control winding of each of the three, three-phase saturable reactors, L3, L4, and L5 (fig. 16-12). This serves as the quiescent control winding input. The 10-11 (power) winding of L1 (fig. 16-11) is placed in the output conduction path so that the error voltage is introduced in series with the secondary voltage of T3. The amount of the output control current at terminals 2 and 3 of TB1 is therefore a function of the magnitude and phase of the error voltage input relative to the T3 input.

The variable current output at the 10-11 winding of the saturable reactor, L1, when error signals are received, is applied via the 3-2 input terminal (fig. 16-12) to the power amplifier control winding. The power winding of each reactor is connected in series with one phase of the line power. The circuit of the bottom reactor is from one side, C, of the 208-volt, 3Ø input, through CR33, the feedback winding of L5 (in parallel with R12), CR36, and the power winding of L5, and through terminals 2 and 3 of T4 (discussed later), to return to A at the 3Ø output terminal. Each of the other power amplifier saturable reactors (L3 and L4) has its feedback and power windings similarly connected in series with one phase of the three-phase power line through a bridge rectifier. Thus, the line current to the spin motor is controlled by the impedance of the L3, L4, and L5 reactors.

For low values of d-c control current, the saturation of the reactors (L3, L4, and L5) is low, and the impedance is high. In this condition, the voltage applied to the motor, B4 (fig. 16-3) is decreased, and the motor will slow down. Conversely, for large values of d-c current, the motor, B4, voltage will be increased due to the lower reactor impedance, and the motor speed increases.

A bias rectifier (comprising CR21 through CR24), which is connected across one phase of the input, supplies a d-c current to the bias windings of L3, L4, and L5, through limiting resistors, R13, R14, and R15, respectively. Each bias winding is shunted by a single rectifier, which is connected so that the rectifier is normally nonconducting. The rectifiers, CR37, CR38, and CR39, respectively, will conduct for spurious voltages, thereby suppressing harmonics of the line frequency that are introduced in the control and bias windings of the reactors.

To correct the power factor of the spin motor voltage, three power factor correction circuits in a wye connection are provided. These circuits include T4 and C7, T5 and C8, and T6 and C9, respectively. Ordinarily, only capacitors would be required, but since the amount of capacity necessary would make the equipment too large, an autotransformer is employed. This autotransformer effectively multiplies the capacity by the turns ratio of the autotransformer squared. This in effect produces a larger capacity across the respective phases of the input power.

RADIO SET AN/SRN-6

Radio Set, AN/SRN-6 (fig. 16-13) is representative of the more recent developments of shipboard tacan equipments. The AN/SRN-6 and

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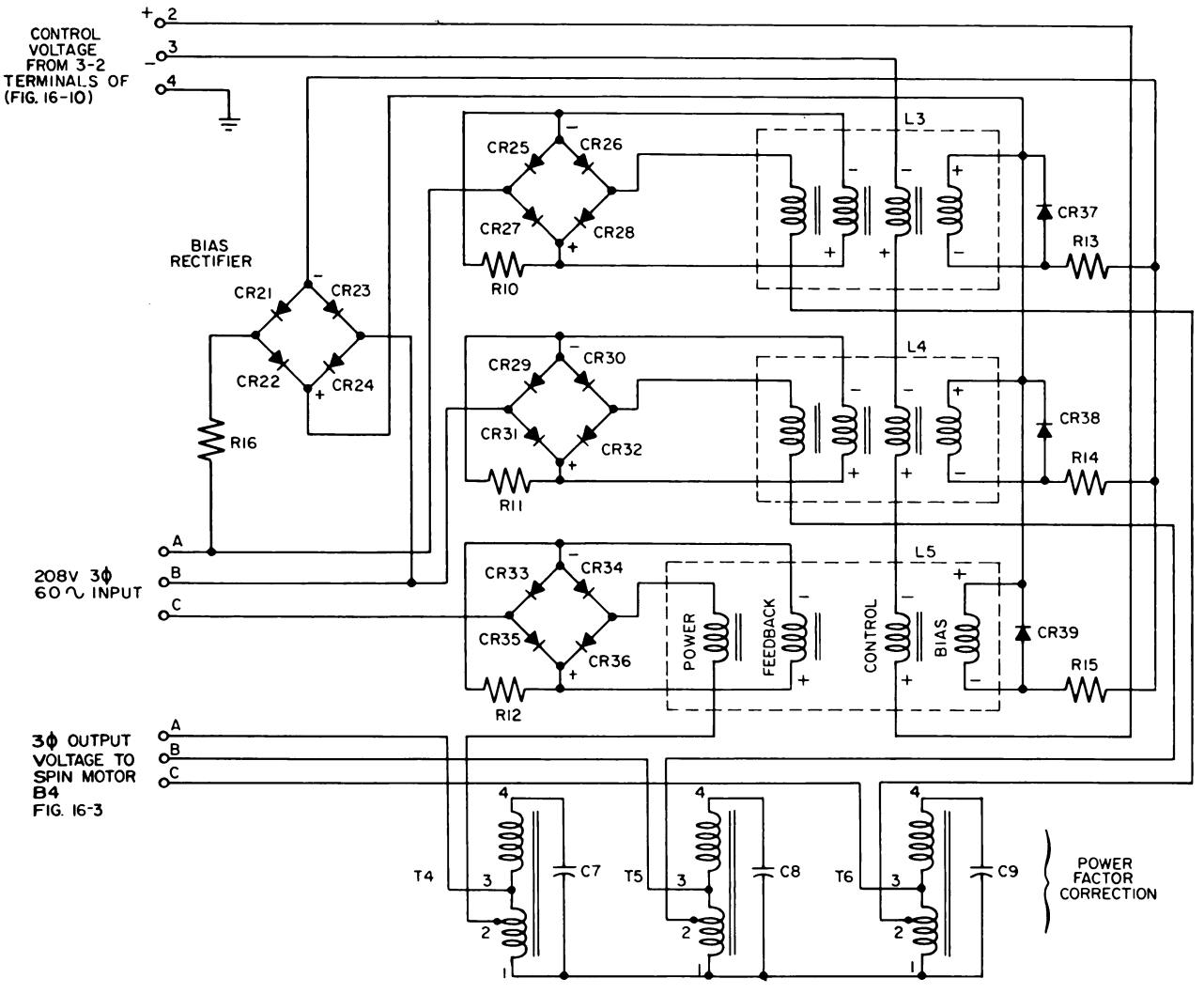

Figure 16-12.—Antenna speed control power amplifier circuit.

AN/URN-3 radio sets are very similar and capable of transmitting and receiving the same type signals. Actually, very few AN/SRN-6 navigational systems are installed aboard ship at this time, although it is anticipated that the AN/SRN-6 equipment will completely replace the AN/URN-3 tacan systems within a two-year period. For this reason, the major differences in the two tacan systems (AN/URN-3 and AN/SRN-6) will be considered in the remaining portion of this chapter.

The information supplied to the aircraft from the AN/SRN-6 is the same as that originating from the AN/URN-3 Radio Set, namely: (1) the direction of the beacon with respect to the aircraft (as referenced to magnetic north); (2) the radio beacon identification code; and (3) the range or distance of the aircraft from the beacon.

The major components of the two tacan systems are essentially the same. They differ mainly in the transmitter output power, the circuit construction of some of the stages of the frequency multiplier-oscillator, and coder-indicator, the type of antenna group used, the primary power supply requirements, and the power distribution circuitry. One major improvement in the tacan system, which is incorporated in the AN/SRN-6, is the addition of built-in test equipment.

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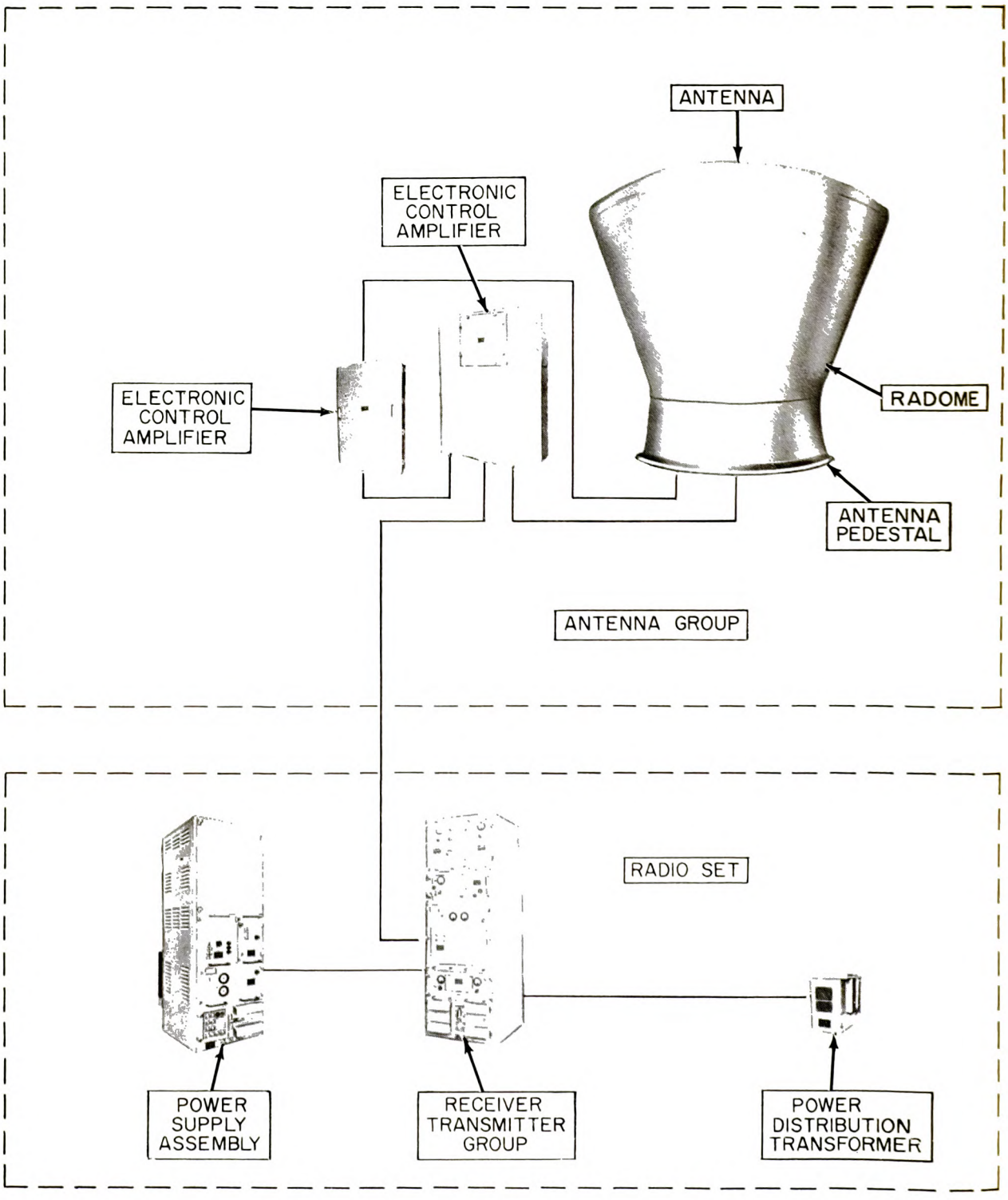

Figure 16-13.—Shipboard Radio Beacon, AN/SRN-6.

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The test equipment is located in the upper section of the power supply assembly of the AN/SRN-6 radio beacon (fig. 16-14). Proper use of these test units enables the technician and/or the operator of the tacan system to ascertain the salient condition of the equipment. The test units included are:

(1) Pulse Analyzer-Signal Generator, TS-890/URN-3
(2) Pulse Sweep Generator, SG-121/URN-3
(3) Oscilloscope, OS-54/URN-3
(4) Power Meter-Pulse Counter, TS-891/URN-3
(5) Switch-Test Adapter, SA-420/URN-3

The Switch-Test Adapter, SA-420/URN-3, is mounted externally, and is not included in the diagram.

A brief description of all the test equipments will be given toward the end of this chapter.

RADIO RECEIVER

The radio receiver of the AN/SRN-6 functions in essentially the same manner as that described for the receiver of the AN/URN-3, and does not require further consideration in this discussion.

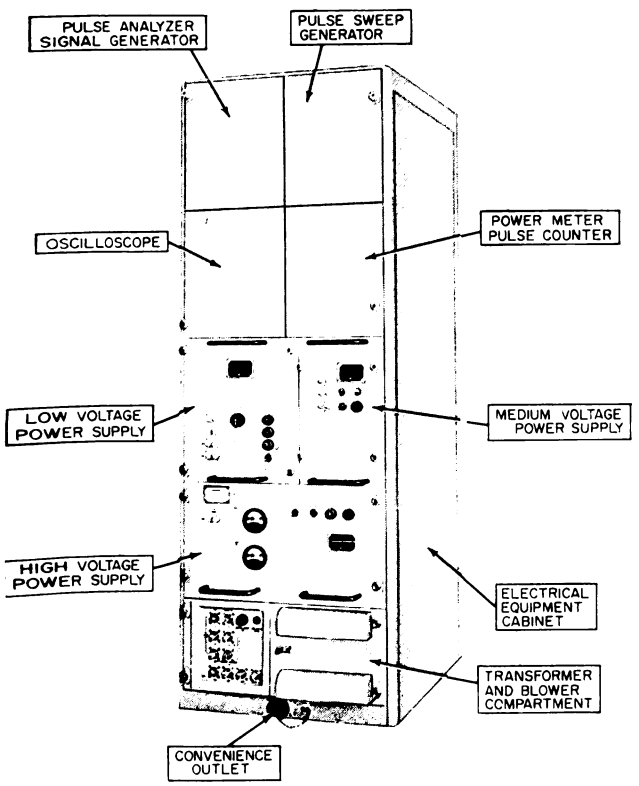

Figure 16-14.—Power supply test group.

CODER-INDICATOR

The coder-indicator (fig. 16-15) of the AN/SRN-6 and the AN/URN-3 produces multiplexed output signals consisting of three components. These are: (1) the reference burst, (2) the beacon identification call tone pulses, and (3) the distance reply pulses. The generation of the latter two signals is almost identical in both tacan systems.

The major differences that exist in the coder-indicator of the AN/SRN-6 as compared to the AN/URN-3 are in the method by which the 135- and 15-cps reference burst pulses are produced. In the AN/URN-3 the burst oscillators (V4 and V9 in chapter 15, figure 15-8) operate at 166.667 and 133.33 kc, respectively, to produce the basic reference burst cycles. The oscillator output frequency is divided in two frequency divider stages, V21 and V22. The resultant reference burst pulses are shaped and amplified in the subsequent reference burst stages. This process is explained in chapters 14 and 15 of this training course.

In the AN/SRN-6, the 135- and 15-cps trigger pulses are derived from the antenna pulser plate (chapter 14, figure 14-5, B) in the same manner as for the AN/URN-3. These pulses are amplified in V1 and V6 (fig. 16-16) and used as the input trigger pulses to the 135- and 15-cps reference burst generators, V3 and V8, respectively. The gate generator circuits are synchronized "one-shot" multivibrators whose negative square wave output is used to trigger its associated ringing oscillator circuit, V4 or V9. The ringing oscillator, in turn, produces the required number of reference burst cycles. These pulses are amplified in V16 and applied to subsequent stages (in the coder-indicator) which are similar to those discussed in chapters 14 and 15.

The positive output from the gate generators, V3 and V8, is applied to the priority gate stage, V15, through a mixing network. The priority gate stage, V15, functions in a similar manner to the reference burst gating amplifier (see V15, chapter 15, figure 15-8) to assign first priority to the reference burst output. The identification code is second in priority, with the distance reply pulses (interrogation signals) occurring last in the priority arrangement.

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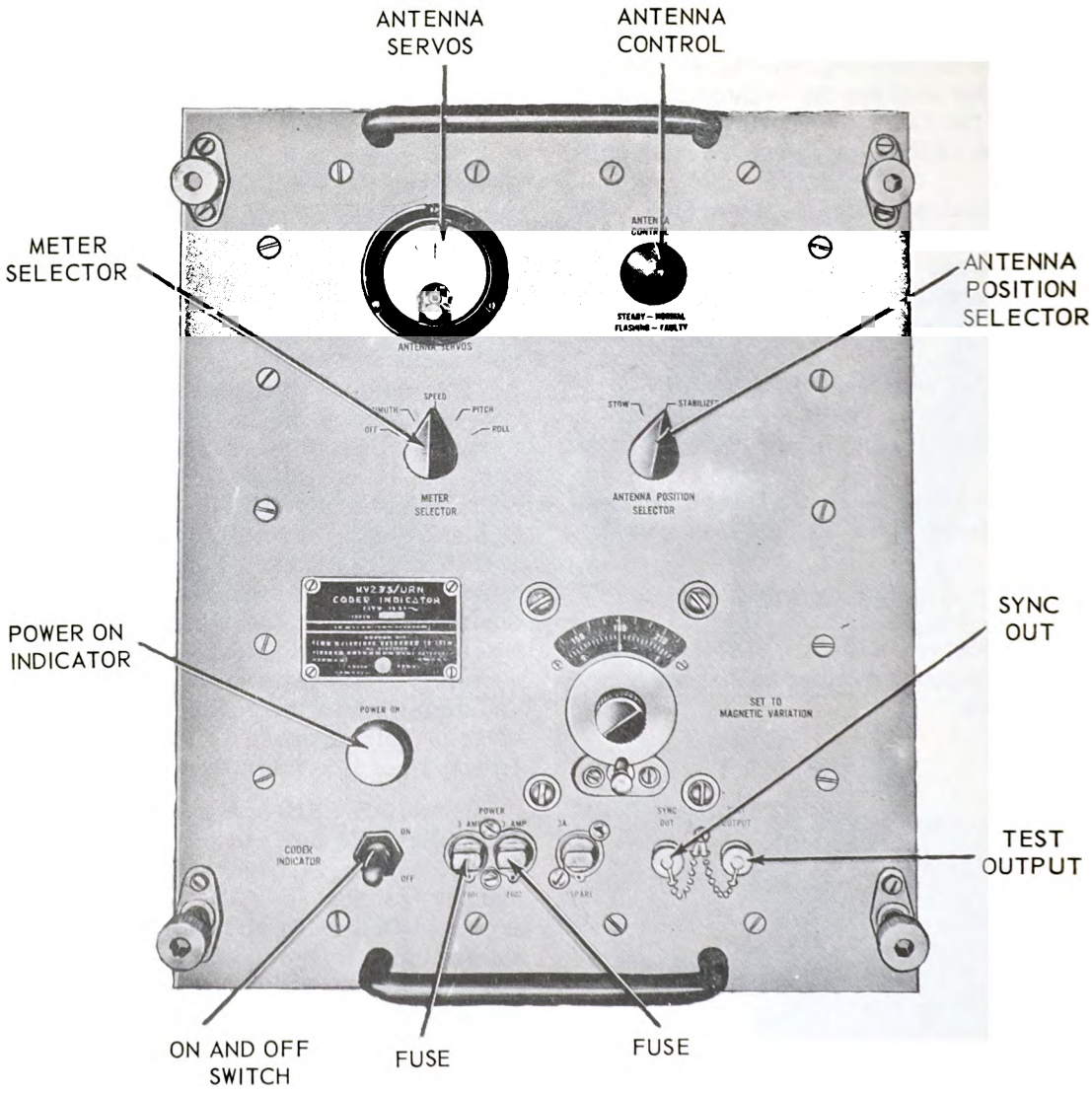

Figure 16-15.—Coder indicator chassis and front panel controls.

The mixing network serves as an isolating and coupling circuit between the two reference burst generators.

Synchronizing pulses are applied to the 135- and 15-cps reference burst gate generators, V3 and V8, to ensure that the gate pulse to the ringing oscillator is just long enough for the oscillator to produce the required number of reference burst cycles.

FREQUENCY MULTIPLIER-OSCILLATOR

The frequency multiplier-oscillator chassis (fig. 16-17) performs the following five major functions: (1) generates the fundamental frequency of the tacan system by means of a crystal oscillator, and multiplies this frequency until it is that of the channel frequencies; (2) provides the radio receiver with a local oscillator signal; (3) provides r-f drive power to the klystron; (4) generates a 10-μs square wave; and (5) generates a 3.5-μs shaped pulse.

Note that a single oscillator (see V1, chapter 14, figure 14-11) in the AN/URN-3 generates the basic frequency for the receiver local oscillator, and the high- and low-band frequencies of the beacon transmitter. In the

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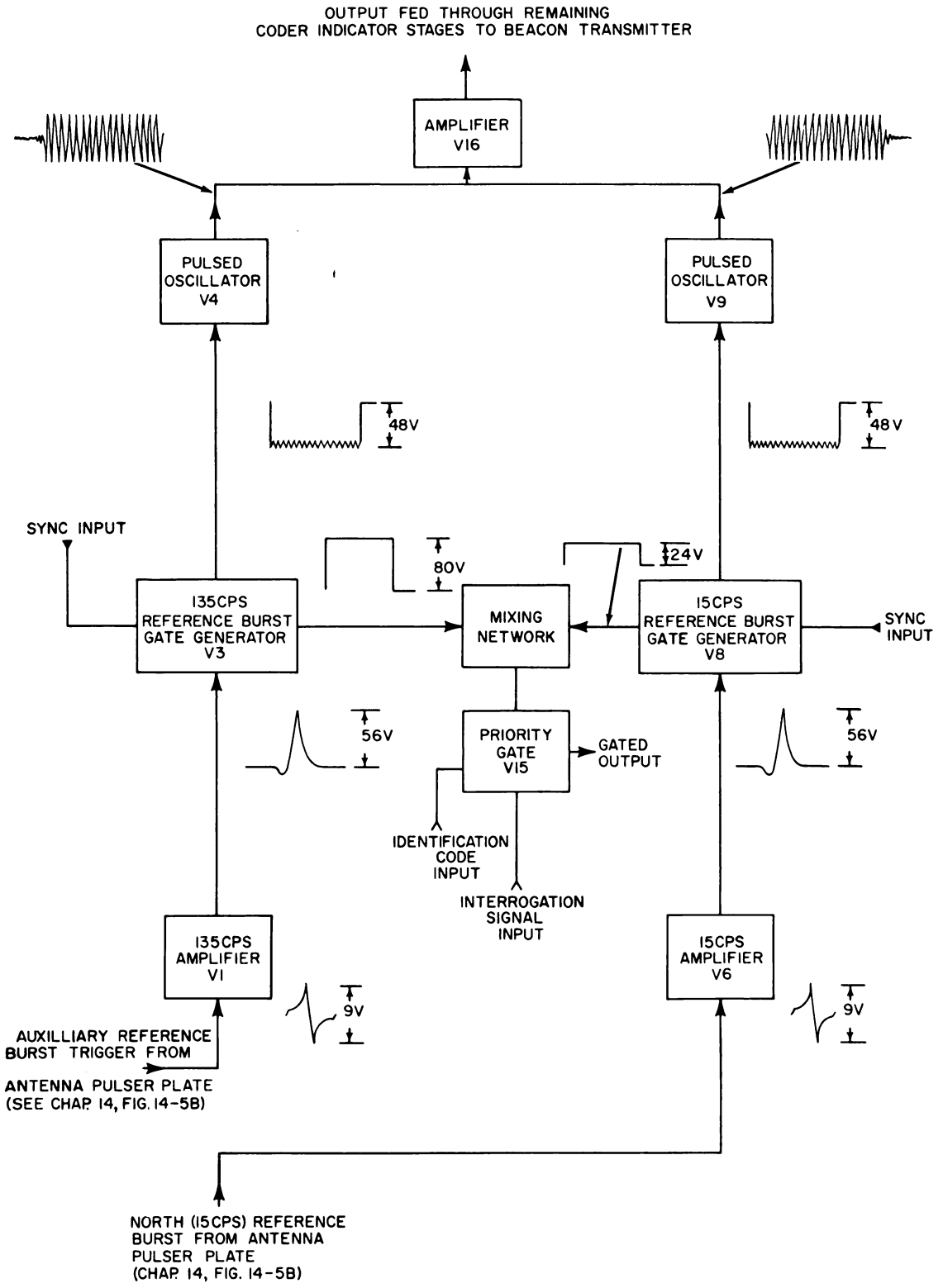

Figure 16-16.—Reference burst generator, block diagram.

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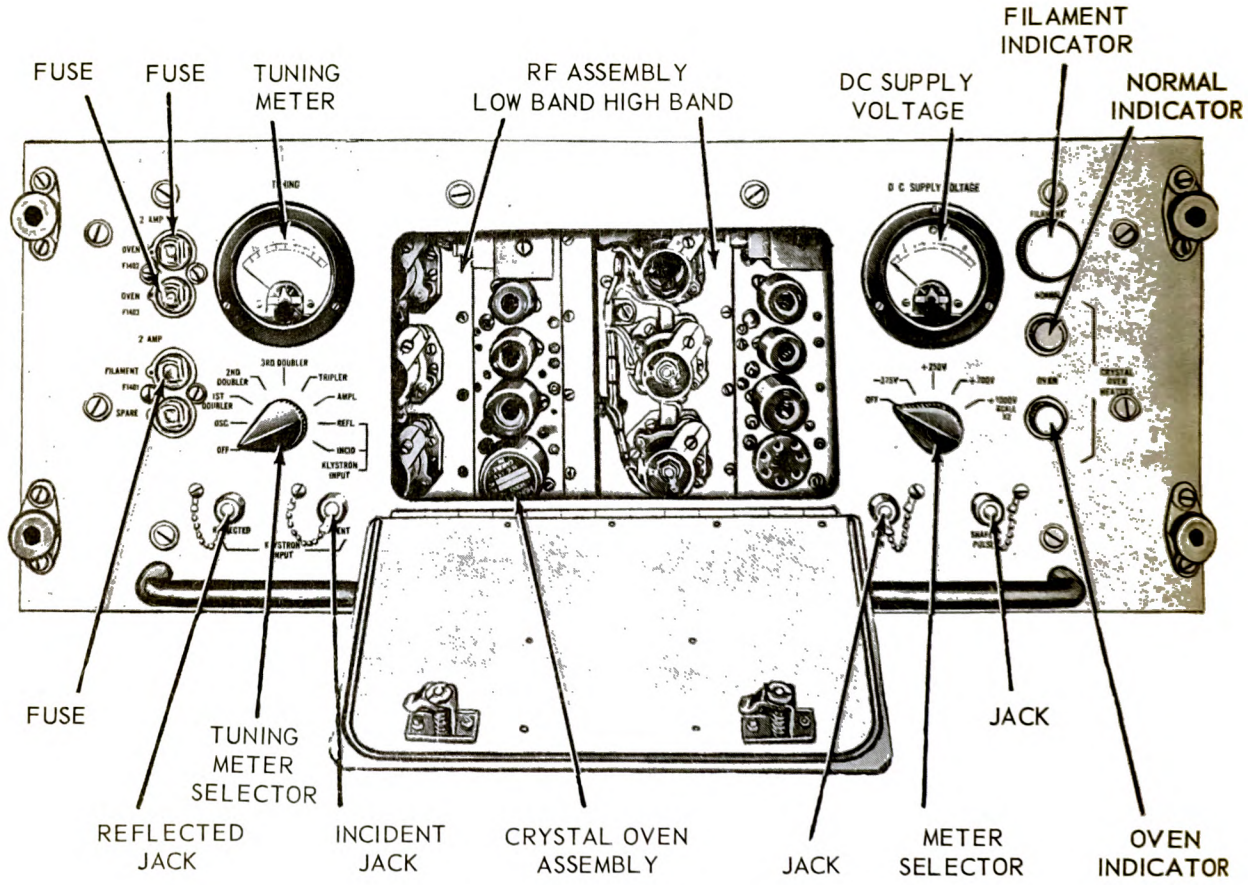

Figure 16-17.—Chassis and front panel controls.

AN/SRN-6 tacan system, two complete crystal oscillator and frequency multiplier chains are provided. This group of stages comprises a unit referred to as the exciter r-f section. One exciter r-f section is used for generating frequencies in the range from 962 to 1,025 mc (low-band operation), and the other is used for generating frequencies between 1151 and 1213 mc (high-band operation). Except for frequency range, the following description applies to both exciter r-f sections.

The oscillator frequency in each of the exciter r-f sections is multiplied in three doubler and one tripler stages (fig. 16-18). The tripler of the respective exciter supplies the local oscillator signal to the beacon receiver.

Another portion of the tripler output is constantly applied to the input of the first amplifier, V5. The output of V5 is applied to V6.

You will recall that the r-f outputs from the first and final doubler stages (V3 and V6, chapter 14, figure 14-11) in the frequency multiplier-oscillator chassis of the AN/URN-3 are keyed by 7.0 and 3.5-μs pulses, respectively. In the AN/SRN-6, the keying pulses are applied to the first and second amplifiers (V5 and V6 of figure 16-18) and to the klystron, V15. The duration of the pulses is 10 us and 3.5 μs, respectively. Keying the stages of the beacon transmitter diminishes the physical size of the succeeding amplifiers and of the power supplies by reducing the duty cycle of the r-f stages.

In the frequency multiplier-oscillator of the AN/SRN-6 tacan system, type 2C39A lighthouse tubes replace the type 4X150G tubes used in the frequency multiplier-oscillator stages of the AN/URN-3. The 2C39A and 4X150G tubes operate similarly into coaxial type resonant cavities.

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In the frequency multiplier-oscillator of the AN/SRN-6, the first and second keyed r-f amplifiers, V5 and V6, are biased beyond cut-off. A 10-μs pulse is generated in a one-shot multivibrator, V7 (fig. 16-18), amplified in V8, and applied as a negative-going trigger (keying) pulse to the cathodes of the first and second amplifiers, V5 and V6, respectively. This pulse gates or keys the first and second amplifiers, thereby permitting the passage of the tripler r-f output to the klystron power amplifier, V15.

A 3.5-μs pulse is generated in the frequency multiplier-oscillator by the stages comprising V9 through V12. The circuits that generate these pulses are similar to the stages that generate the 3.5-μs pulse in the AN/URN-3. In both systems, the 3-5-μs pulse is used to intensity-modulate the klystron (V15) beam current.

AMPLIFIER-MODULATOR

The amplifier-modulator (fig. 16-18) consists of an r-f power amplifier, V15, a regulated bias power supply, and a double-slug tuned circuit, Z1, which provides a means of matching the output impedance of the klystron with the input impedance to the control-duplexer.

The r-f power amplifier, V15 of the AN/SRN-6, employs a SAL-89 type klystron as opposed to SAL-39 used in the AN/URN-3. The minimum peak power output of the SAL-89 is 7.5 kw, while the minimum peak power output of the SAL-39 is 5 kw.

The klystron is a three-cavity amplifying tube, which has a control grid used to intensity-modulate the klystron current, and an r-f input jack connected into the first cavity, which is used to velocity-modulate the klystron current.

A constant negative potential of 12 kv is applied to the klystron (V15) cathode from the high voltage power supply. In the quiescent condition (no klystron input), the klystron beam current is cut off by approximately 120 volts of negative bias obtained from the regulated bias supply in the amplifier-modulator unit.

The action of the klystron power amplifier, V15, is similar to a coincidence stage. V15 receives the 10-μs keyed r-f input and the 3.5-μs pulse. The 10-μs pulse is applied to the first cavity of the klystron to velocity-modulate the klystron beam current. The 3-5-μs pulse is applied to the control grid of the klystron during the center portion of the 10-μs pulse to intensity-modulate the klystron beam current. The two simultaneous actions produce a 3.5-μs r-f pulse envelope.

The 3.5-μs r-f pulse from the klystron, V15, has a minimum peak power of 7.5 kw and a repetition rate identical to the repetition rate of the pulse train at the output of the coder-indicator. The output of the coder-indicator is one second for the AN/URN-3 and AN/SRN-6 consists of 3600 pulse-pairs with a 12-μs spacing between pulses of a pair.

CONTROL-DUPLEXER

The control-duplexer is a passive network, which permits both the transmitter and receiver to be operated from the same antenna. In the control-duplexer of the AN/URN-3 (fig. 16-19, A) a variable length open-circuited stub, Z2, is connected across the output transmission line. This stub is an integral (whole) number of half wave lengths long at the transmitter frequency so that it acts as an open circuit at the transmitter frequency. At the receiver frequency, the open circuited stub is an odd number of quarter wave-lengths long so that it acts like a short circuit. Thus, the signal from the transmitter is allowed to pass, while signals from the antenna are short circuited at the transmitter output.

In the AN/SRN-6, the pulsed r-f transmitter output from the klystron r-f amplifier, V15 (fig. 16-18) is fed to a transmission line filter (Z2 or Z3 of figure 16-19, B), which consists of a pair of tunable resonant cavities. The transmission line filter is located in the line between the transmitter and the antenna, with the receiver input connection branching off on the antenna side of the filter. As the transmission line filters, Z2 or Z3, are tuned 63 mc above or below the receiver frequency, Z4 appears as an open circuit at the transmitter frequency. Thus, the transmitter output is fed directly to the antenna.

Because of the high Q's associated with resonant cavities, the response curve of the control-duplexer in the AN/SRN-6 will be sharper than the curve for the control-duplexer of the AN/URN-3. In this way, the transmission of adjacent frequencies is reduced in the AN/SRN-6.

In the control-duplexer of the AN/SRN-6 (fig. 16-19, B), Z2 is the low-band filter used at frequencies between 960 and 1025 mc. Z3 is the high-band filter, and is used at frequencies between 1150 and 1215 mc. The performance

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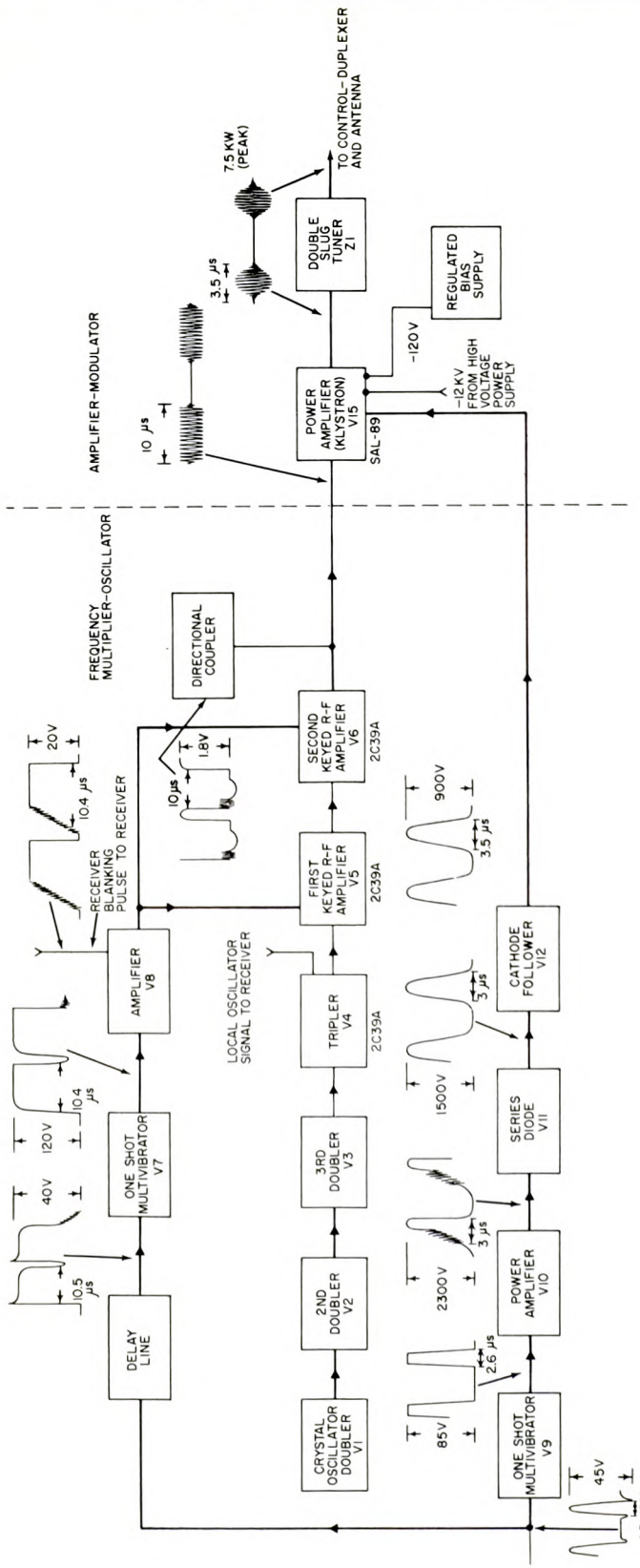

Figure 16-18.—Frequency multiplier-oscillator and amplifier-modulator, block diagram.

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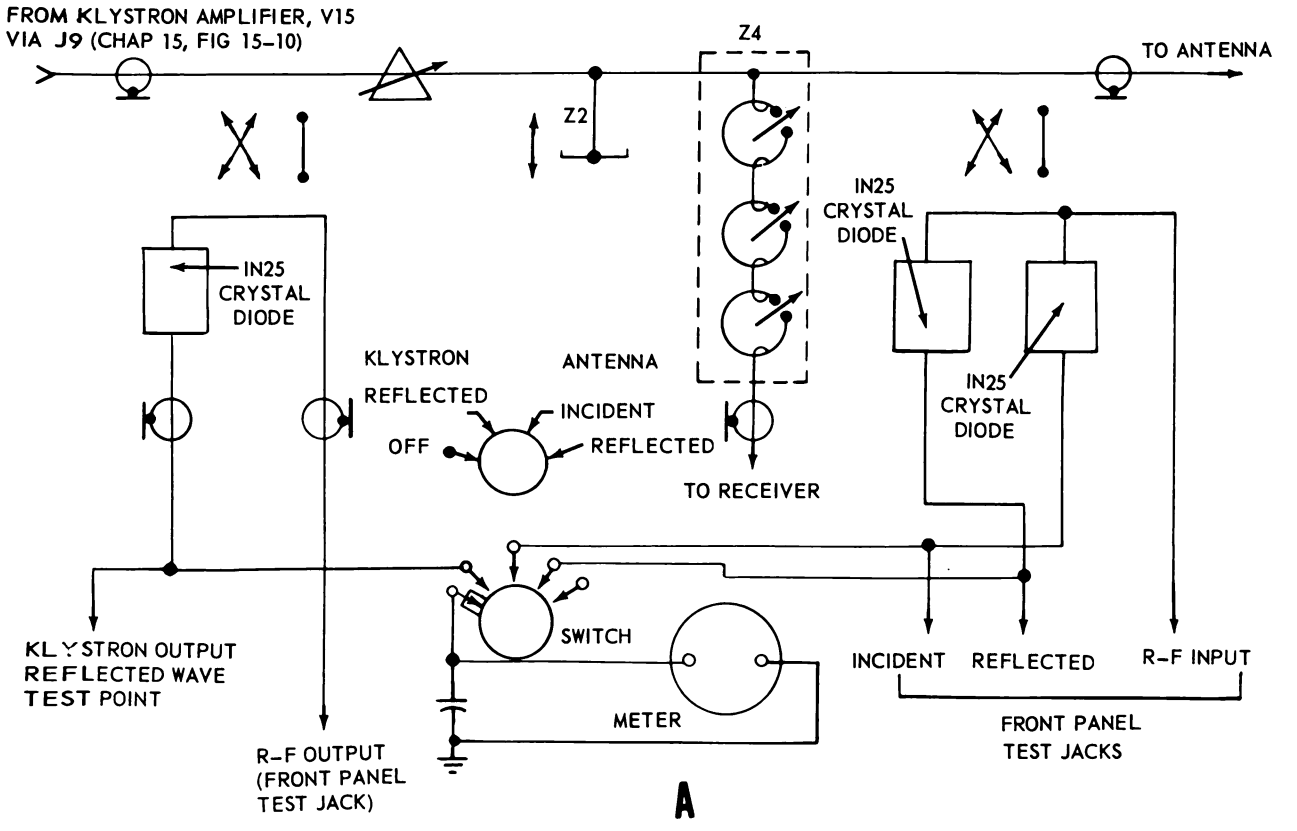

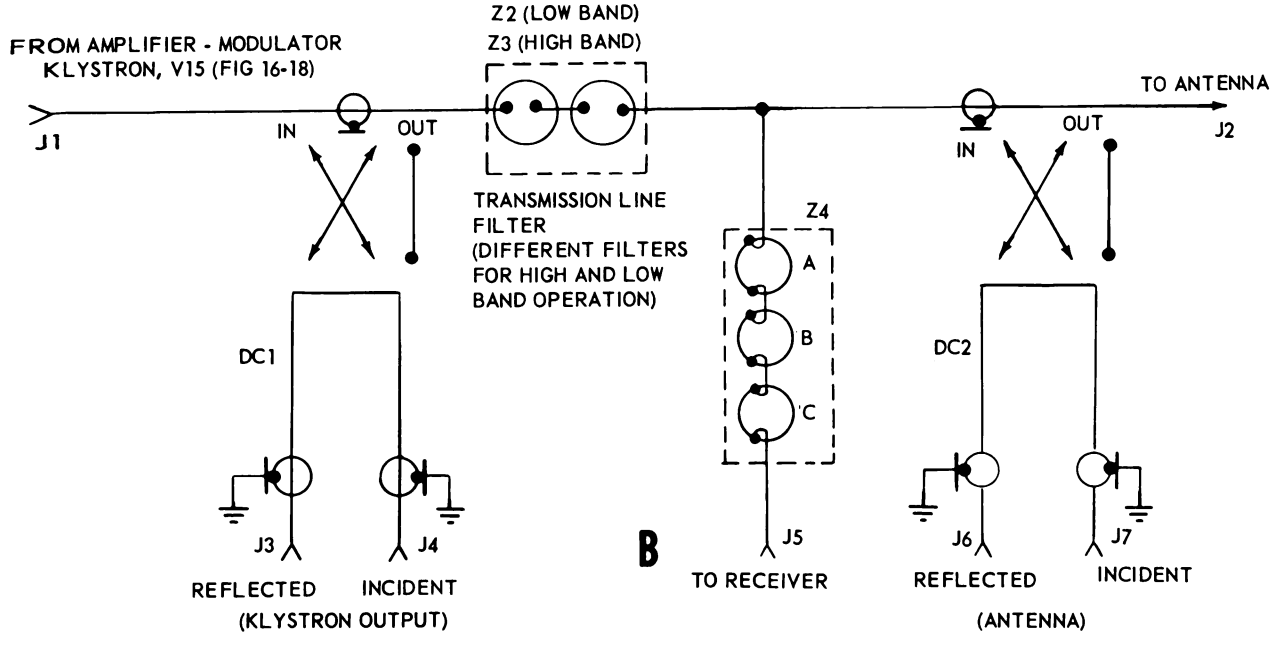

Figure 16-19.—Control duplexer.

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standards for the transmission line filters are as follows:

1. The signal insertion loss at the transmitter frequency is about 1.4 db maximum.
2. Within ± 2 mc of the transmitter frequency, the filter response is within 3 db of maximum.
3. At ± 0.75 mc from the transmitter output frequency, the filter response is a minimum of 20 db down.
4. The transmitter response curve is down sufficiently at the receiver frequency (transmitter frequency ± 63 mc) so that the transmitter frequency is completely rejected at the receiver.
5. The temperature operating range of the filter is from -54° to +65° C.

Spaced one-half wavelength from the output terminal of the transmission line filter is a Tee connection to which is attached the receiver preselector cavities comprised in Z4. The operation of the filter, Z4, in the AN/URN-3 and AN/SRN-6 is the same. The preselector cavities, A, B, and C, of Z4 are tuned to the receiver frequency, which is 63 mc above the transmitter frequency for the low-band, and 63 mc below the transmitter frequency for high-band operation. Because of this wide frequency separation, signals from the antenna at the receiver frequency are routed to the radio receiver and signals at the transmitter frequency are rejected.

Samples of the klystron-incident and klystron-reflected voltages and of the antenna-incident and antenna-reflected voltages are made available by directional couplers, DC1 and DC2, respectively. During tuning, the klystron output may be obtained at J3 and J4, while the antenna output can be obtained at J6 and J7.

The control-duplexer contains a desiccator (drying agent) that controls the humidity of the air in the two-filter cavities (Z2 or Z3 and Z4). Due to temperature changes, the filter cavities tend to breath (draw in and expel air). This air exchange takes place through the desiccator, with the desiccant absorbing moisture in the air, thereby maintaining a relatively dry atmosphere in the cavities.

POWER SUPPLIES

No differences exist in the low-voltage power supplies of the AN/URN-3 and AN/SRN-6 tacan systems that are sufficient to warrant discussion here.

Automatic protection against sustained overloads is provided within the medium-voltage power supply of the AN/SRN-6. In the AN/URN-3, a time delay, or lag, in the overload circuit prevents activation of the overload relay during the period of heavy charging currents when the equipment is initially turned on.

To establish whether or not a sustained overload exists in the AN/SRN-6, the protective circuit turns the power supply off and on three times in succession before turning it off completely. The overload circuit is connected into both the medium- and high-voltage power supply circuits. Thus, an overload condition affecting either one of the two power supply units will deactivate both power sources.

The high-voltage power supply of the AN/SRN-6 furnishes -12,000 volts unregulated direct current to the klystron amplifier (V15 of figure 16-18). This is in contrast to a +12-kv potential, which is used in the klystron beam stages of the AN/URN-3.

The input to the high-voltage power supply is 9220 volts, three-phase, 60 cps alternating current. The power supply rectifiers are full-wave circuits operated in series-parallel. A single capacitor filter is employed.

ANTENNA GROUPS

The shipboard antenna groups for the AN/SRN-6 tacan equipment are the OA-1545/SRN-6 for low-band operation and the OA-1546/SRN-6 for high-band operation. These antenna groups (fig. 16-13) are essentially the same as those used with the AN/URN-3 except for the antenna speed control tachometer voltage, the antenna gain, and the number of coaxial sections that make up the central array.

The antenna gain for the AN/SRN-6 beacon is about 4.7 db, which is 1-3 db less than the 6 db gain of the AN/URN-3 antenna. However, with minor alterations in the circuit connections, the antennas for both systems can be made interchangeable.

The central array of the OA-1545/SRN-6 and the OA-1546/SRN-6 contains four vertical biconical dipole sections as opposed to 7 sections for the AN/URN-3 antenna (see chapter 14, figure 14-17). The frequency range of the two antennas is the same.

The antenna speed control tachometer of the AN/SRN-6 (not shown) supplies a 675-cps signal to the antenna speed control circuits when the antenna cylinders are being rotated at 900 rpm.

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You may recall that the frequency of the tachometer feedback in the AN/URN-3 is 810 cps when the antenna speed is correct (see figure 16-10 and 16-11). Although the frequency of the feedback is different for the two systems, the control circuits are basically the same.

A new antenna group (OA-1801/SRN-6), which has about the same gain as the AN/SRN-6 antennas just discussed, has recently become a part of the AN/SRN-6 tacan system. This antenna, however, is much lighter than the previous antennas (295 lbs as opposed to 750 lbs) and is therefore much easier to drive at the required 900 rpm. Thus, fewer breakdowns result because of the defects in the belt arrangement that drives the antenna.

The central array of the antenna group, OA-1801/SRN-6, has only two vertical dipoles. The gain of this antenna is still maintained at 4.7 db.

BUILT-IN TEST EQUIPMENT

Built-in test equipments are provided to permit rapid checking of the important operating characteristics of the radio beacon. The cabinet that contains most of the test equipment is shown in figure 16-14. A brief description of each of the test units is given below.

Pulse-Sweep Generator, SG-121A/URN-3

The pulse-sweep generator (fig. 16-20) provides pulse-pairs at a controlled variable rate to simulate the pulse output of one or more airborne radio sets, AN/ARN-21. The pulse-sweep generator pulse output is not applied to the tacan receiver, but is first fed to a pulse-analyzer-signal generator (TS-890A/URN-3 discussed later) and used in this unit to produce the required pulsed r-f signal, which is similar to the normal output of the airborne radio set. This pulsed r-f voltage is then applied to the radio beacon receiver.

The paired-pulse output of the Pulse-Sweep Generator, SG-121A/URN-3, is carefully timed. The spacing between pulses in a pair may be varied from 11 to 13 μs in 0.5-μs steps. This permits checking the radio beacon output pulse-pairs to ensure that the space between pulses of a pair is within the specified limit of 11.5 to 12.5 μs and that the beacon will decode and retransmit only those pulse-pairs with time spacing within this limit.

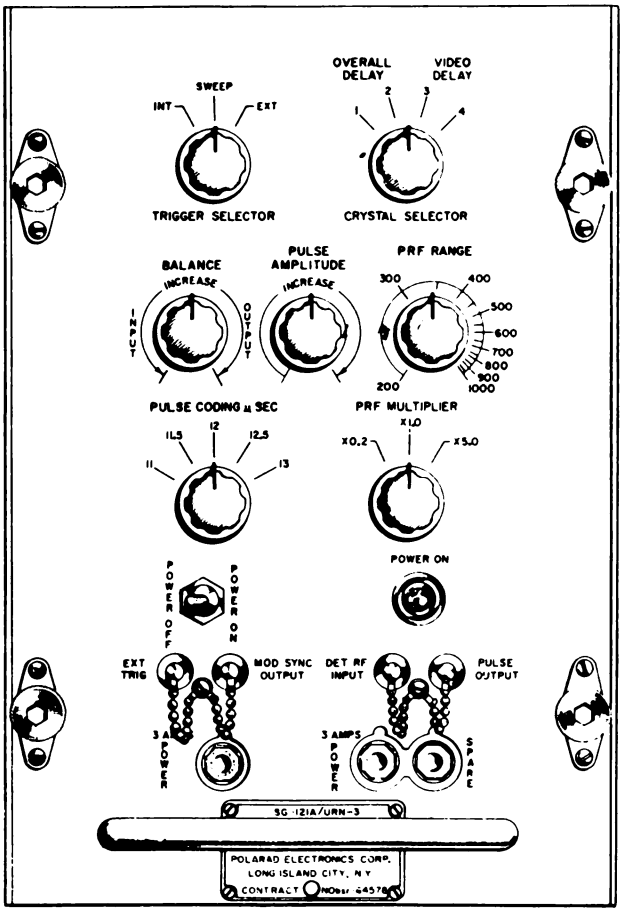

Figure 16-20.—Pulse-Sweep Generator, SG-121A/URN-3, front panel.

The pulse repetition rate of the pulse-sweep generator output is continuously variable over a range from 40 to 4000 pulse-pairs per second. This permits the pulse-pairs per second at the input to the beacon receiver to be varied, thereby interrogating the beacon at rates that simulate different numbers of radio sets (AN/ARN-21) within the operating range. Fixed crystal-controlled rates are also provided.

Separate crystal timing frequencies produce sweep rates for checking the upper and lower tolerance limits of the 50-μs "zero distance relay." The limits are 50.2 and 49.8 μs, respectively.

Pulse Analyzer-Signal Generator, TS-890A/URN-3

The pulse analyzer-signal generator consists of a signal generator section, which provides an

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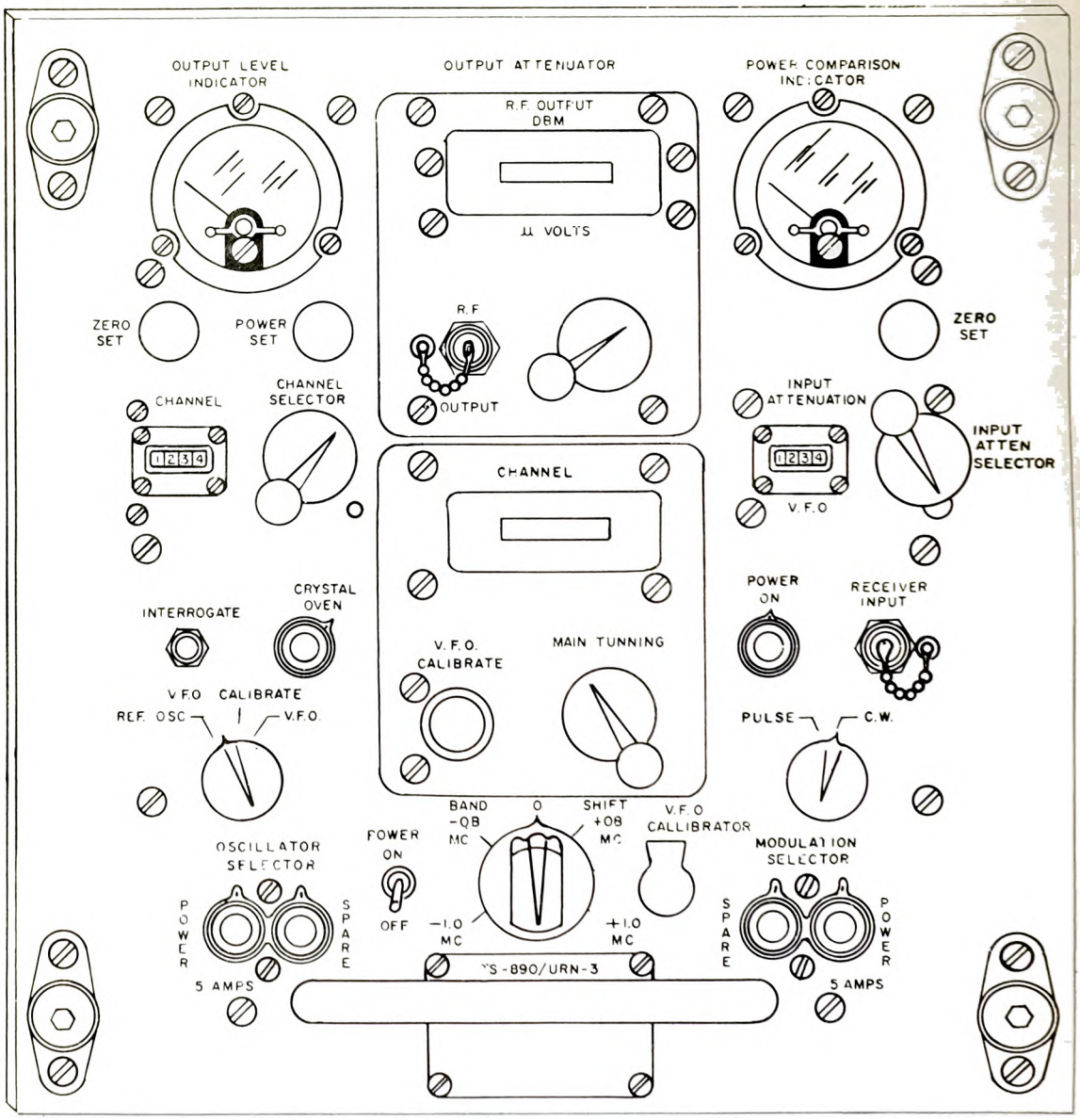

Figure 16-21.—Pulse Analyzer-Signal Generator, TS-890/URN-3, front panel.

r-f carrier for testing the beacon, and a calibrated receiver (pulse-analyzer), which is used for making spectrum analysis of the beacon output signal. The pulse analyzer-signal generator front panel is shown in figure 16-21. The r-f signal provided by the signal generator portion of this unit is continuously variable over the range from 1025 to 1150 mc. Plug-in crystals may be employed for any channel within the 1025- to 1150-mc range.

The amplitude of the signal generator output (when the generator is terminated in 50 ohms) may be varied over an accurately calibrated range of 10 μv to 0.5 volt. The r-f signal output may be either continuous wave or pulsed.

The pulse analyzer (receiver) can be tuned to any frequency in the two bands from 962 to 1025 mc and from 1150 to 1213 mc. The received signal is applied to a metering circuit through a very narrow-band amplifier. The pulse

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nalyzer can be tuned to the exact center fre-uency of the signal, or to 0.8 or 2.0 mc either ide of the center. This permits metering the pectral distribution characteristics of the out-ut pulse transmitted by the beacon.

)scilloscope, OS-54/URN-3

The Oscilloscope, OS-54/URN-3 (fig. 16-22) s of the synchroscope type, requiring a trigger pulse to produce a sweep trace on the cathode-ray tube. The sweep trigger may be from a recurrent source within the scope or from an external source. The sweep speed may be varied over a range from 0.4 to 40,000 μs per inch. The bandwidth of the internal amplifiers of the scope between those frequencies, which are 3 db down, is from 5 cps to 3 mc.

The scope provides timing markers and calibrating voltage to permit accurate measurement of the signal waveform amplitude and

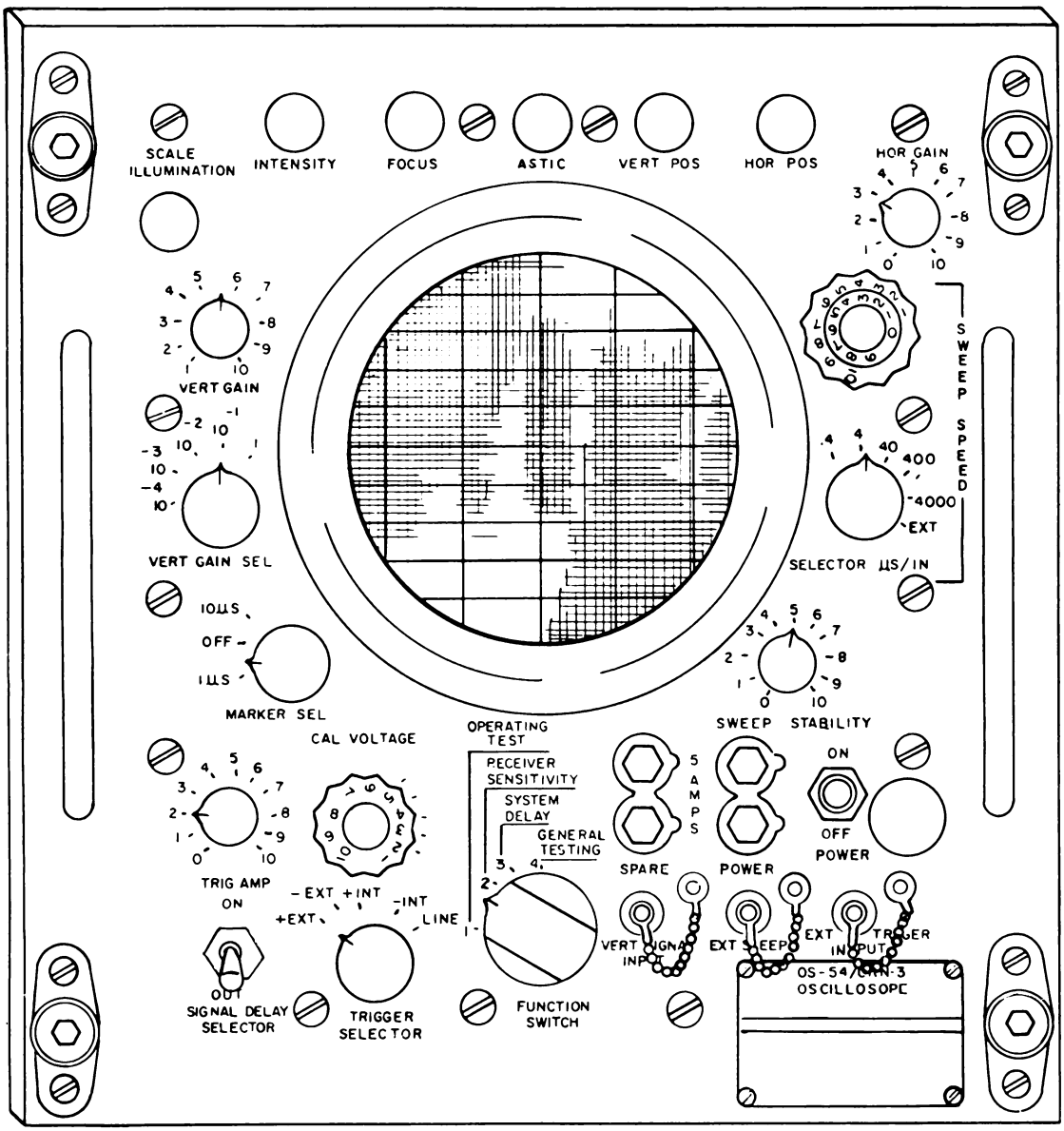

Figure 16-22.—Oscilloscope, OS-54/URN-3, front panel.

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pulse width. The markers are spaced at intervals of 1 to 10 μs, and may be superimposed on the displayed signal. The calibrating voltage permits peak-to-peak amplitude measurements over a range of 0.1 to 100 volts. An optional signal delay of 0.3 μs makes possible the observation of the leading edge of sharply rising waveforms.

Power Meter-Pulse Counter, TS-891/URN-3

The Power Meter-Pulse Counter, TS-891/URN-3 (fig. 16-23) is a single unit, which comprises two functionally separate circuits, the power meter circuit and the pulse counter circuit. The front panel of this unit is shown in figure 16-23.

The power meter circuit is a variable d-c source, which is used to bias a pickoff diode in the Switch-Test Adapter Unit, SA-420/URN-3 (discussed later). The pickoff diode receives a small portion of the beacon transmitter output. Varying the bias on the pickoff diode changes the diode clipping level and causes less of the detected positive signal to be passed on for display on a scope. By adjusting the bias to the point where the detected pulse is just eliminated from the scope presentation, the d-c voltage is made equal to the peak amplitude of the pulse. When the amplitude of the d-c voltage and the pulse peak are made equal, an indication of peak power can be read on a front panel meter. This method of determining the power is known as the "slide-back" method. The meter that indicates the magnitude of the d-c voltage is calibrated in kilowatts.

The pulse counter circuit may perform several functions. When counting the beacon squitter output pulses, the pulse counter accepts these pulses, shapes them for uniformity, and applies them to a vacuum tube measuring circuit, which varies the meter current in direct proportion to the pulse rate. The associated meter is calibrated directly in cps, and has ranges of 80 to 800 and 800 to 8,000 cps. Thus, the meter measures the squitter output directly. Squitter count is the random output of the beacon not due to interrogation.

When counting the reply rate (that is, those pulses that pass through the beacon receiver as a direct result of interrogations) the pulses are applied to the pulse counter in the same manner as for squitter count. Interrogating pulses are applied from the Pulse-Sweep Generator, SG-121A/URN-3 and simultaneously to the counter circuit and to a blocking gate circuit. The gate circuit prevents the passage of interrogating pulses through the counter during a 14-μs period following delivery of the preceding interrogation signal from the pulse-sweep generator.

In addition to the squitter count and reply count, a 4-position function switch (fig. 16-23) permits the counter to accept and count either negative or positive pulses from any external source.

Switch-Test Adapter, SA-420/URN-3

The switch-test adapter (not shown) consists of a heavy duty coaxial switch and a pickoff diode. The switch transfers the radio beacon transmitter output power from the antenna to a dummy load and to the pickoff diode. The diode, in turn, supplies a detected signal to the power meter circuit, as discussed. A biasing voltage developed in the power meter circuit is applied to the diode to permit power measurements by the "slide-back" method.

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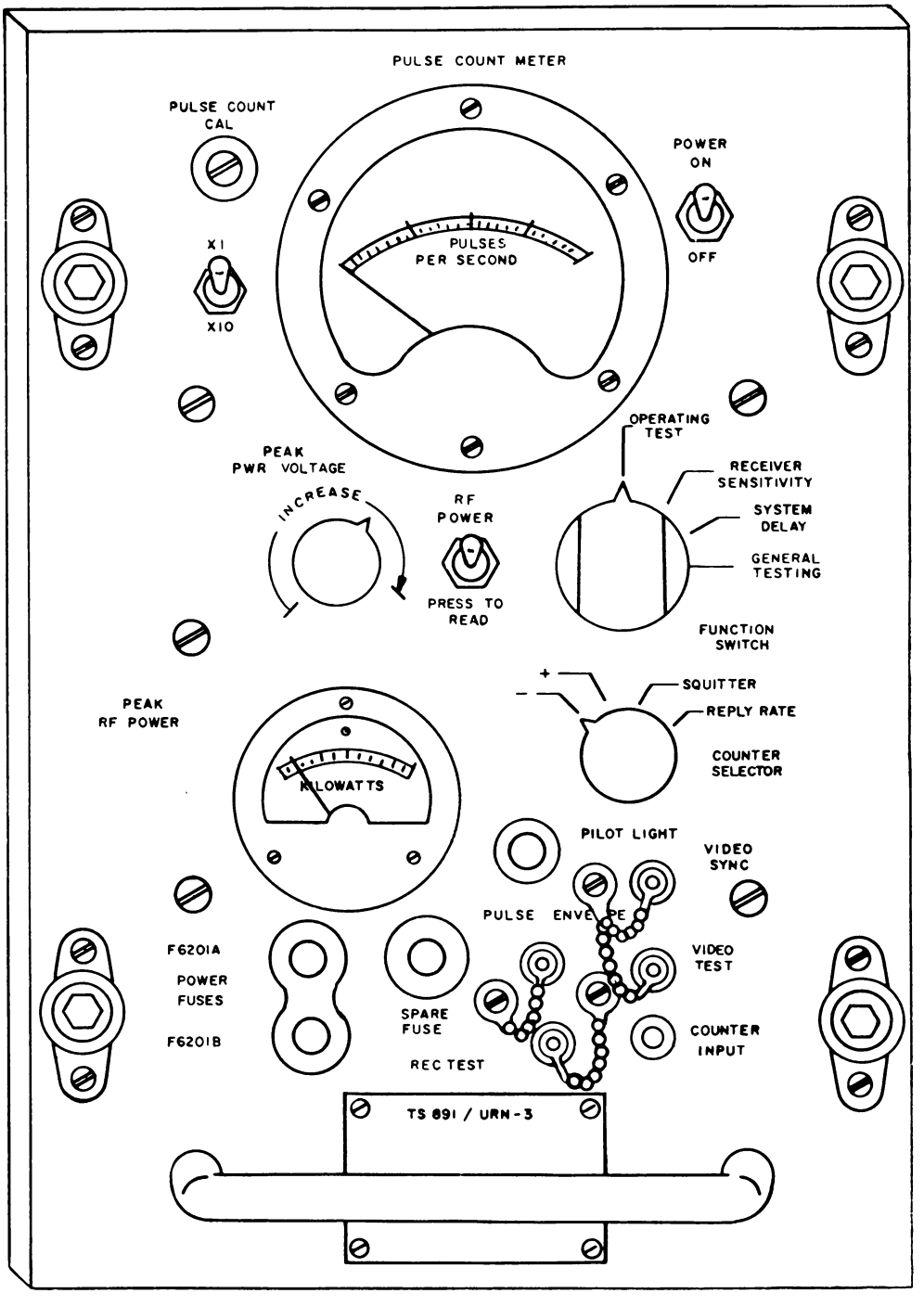

Figure 16-23.—Power Meter-Pulse Counter, TS-891/URN-3, front panel.

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QUIZ

1. The precise sequence and timing of the trigger pulses generated by the reference pulse generator (chapter 14, figure 14-5, B) is determined by what factors?
2. The transmission of the angular position of a mechanical shaft to a remote location electrically may be accomplished by what means?
3. The function of the synchro transmitter (fig. 16-1) is to transmit what type data?
4. The synchro receiver differs from the synchro transmitter in what way?
5. The control transformer control voltage (fig. 16-1) reaches a null (zero volts output) when the control transformer rotor is displaced to what position?
6. The control transformer is used where it is desired to obtain a voltage indicative of what relationship in a servomechanism?
7. The phase of the error signal from the control transformer depends on what factor?
8. The differential transmitter (fig. 16-2) provides a means of adjusting what two components?
9. The magnetic variation signals from the gyrocompass (fig. 16-3) are fed to what units (in shipboard equipments)?
10. The bearing magnetic amplifier (fig. 16-4) detects the phase of the error signal from the control transformer, and amplifies the signal to a level sufficient to drive what unit?
11. To detect a reversal in phase of an a-c signal, it is necessary to compare the signal to what type voltage?
12. The magnitude of the error voltage to the 1- and 36-speed phase detectors (fig. 16-4) is determined by what condition?
13. The fine control (36-speed) transformer voltage (fig. 16-4) is fed to the servo only when the amplitude of the coarse (1-speed) control transformer voltage is below what value?
14. The addition of the antistickoff voltage to the one-speed control transformer voltage (fig. 16-5, B) eliminates what problem in the servo system?
15. A relay, K2 (fig. 16-6) operated by a separate, two-stage magnetic amplifier switches the input of the phase sensitive detector from the 1-speed input to the 36-speed input when the 1-speed input drops below a value corresponding to approximately what value?
16. The switching preamplifier, L10, and switching amplifier, L11 (fig. 16-6) are self-saturating reactors, which are used for what purpose?
17. The preamplifier, L12 (fig. 16-6) is divided into two equal sections, which are connected in what manner?
18. The circuit of the error indicator (fig. 16-6) is what type circuit?
19. Roll and pitch stabilization control facilities are employed in the shipboard radio beacon to compensate for errors introduced into the beacon information by what factors?
20. The isolation impedance in the tachometer circuit (fig. 16-7) serves what other purpose in the circuit?
21. A stabilizing feedback to the booster (fig. 16-9) is obtained by rectifying the a-c voltages, which appear across what windings of the power amplifiers, L4 and L5?
22. What type servo motors are used in the roll and pitch amplifiers?
23. The speed of the rotating cylinders (rotated by the spin motor of figure 16-3) must be maintained at what value?
24. The speed control amplifier circuit (fig. 16-10) receives the tachometer output at a frequency sensitive circuit comprising what components?
25. The power winding of each reactor (fig. 16-12) is connected in what manner with respect to the power line?
26. The information supplied to the aircraft from the AN/URN-3 and AN/SRN-6 Radio Sets is related in what manner?
27. Built-in test equipment is one major improvement that is incorporated in what tacan system?
28. The minimum peak power output of the SAL-89 type klystron employed in the AN/SRN-6 (fig. 16-18) is what value?
29. In the AN/SRN-6, the pulsed r-f transmitter output from the klystron r-f amplifier, V15 (fig. 16-18) is fed to a transmission line filter (Z2 or Z3 of figure 16-19, B), which conists of what type components?
30. The antenna speed control tachometer of the AN/SRN-6 supplies what frequency signal to the antenna speed control circuits when the antenna cylinders are being rotated at 900 rpm?
31. What test equipment provides pulse-pairs at a controlled variable rate to simulate the pulse output of one or more airborne radio sets (AN/ARN-21)?

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CHAPTER 17

THE USE OF SPECIAL TEST EQUIPMENT, PART I

INTRODUCTION

The proper operation of a piece of electronic quipment requires almost constant mainte-ance, either preventive or corrective. To erform the maintenance effectively, it becomes ecessary to use and rely upon various types f test equipment.

It may be necessary on occasion to check the quipment operation to note whether or not it onforms to a standard required for the par-icular equipment.

Test equipment of various types, much too umerous to be listed here, must be at the ommand of the electronic technician. Proper se of the equipment and knowing its capabilities re vital requirements of the user.

The test items considered in this chapter nd chapter 18 of this training course include ie radio interference measuring set, synchro-cope, spectrum analyzer, electronic switch, bsorption wavemeter, and grid dip meter.

RADIO INTERFERENCE MEASURING SET AN/URM-47

The Radio Interference Measuring Set AN/ RM-47 is a highly sensitive h-f, v-h-f, and -h-f superheterodyne receiver covering the equency range from 20 to 400 mc. The equip-ient is used primarily for the location and ieasurement of r-f interference (noise) and)r making field intensity measurements in its equency range.

The total frequency range of the AN/URM-47 i divided into six bands shown in table 17-1. he unit is capable of receiving c-w, m-c-w, ilse-modulated, or radio interference (noise) gnals.

With the proper input impedance, r-f voltage om 1 μv to 1.0 v may be measured. In terms of eld-intensity, (defined later) 1.2 microvolts-er-meter to 60 volts-per-meter (depending on e frequency) may be measured, utilizing the itennas supplied with the equipment. When compared to the outputs from standard signal generators, the accuracy of measurements made by the AN/URM-47 lies within $\pm$15 percent at signal levels above 10 μv.

Table 17-1. — Frequency Bands of the AN/URM-47.

Bands	Megacycles
1	20 to 33
2	33 to 54
3	54 to 88
4	88 to 145
5	145 to 250
6	250 to 400

The major components of the radio interference measuring set (fig. 17-1) include a radio interference field intensity meter, a power supply, and several antennas (not shown) for interference or signal pickup. Either a dipole or loop antenna is used with the equipment.

BLOCK DIAGRAM

A block diagram of the radio interference field intensity meter (hereafter referred to as the RI-FI meter) is shown in figure 17-2. For the sake of simplicity, switches and several controls have been omitted. The purpose and operation of all controls will be considered where their functions are pertinent to the discussion.

The signal channel through the RI-FI meter closely resembles that of a superheterodyne receiver, particularly in its r-f, i-f, and a-f circuits. It differs, however, in its provisions for r-f attenuation and detector output measurements.

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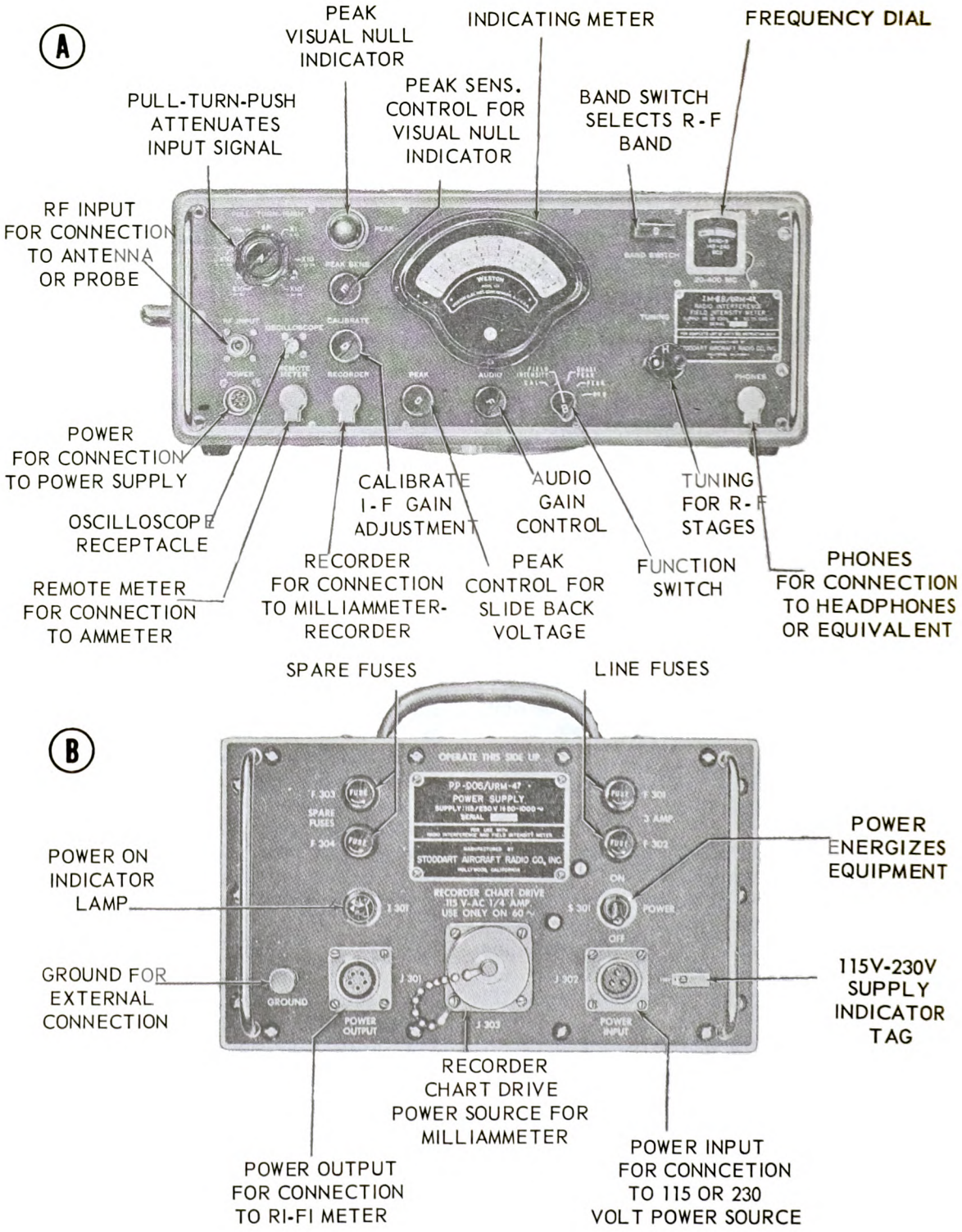
A
PEAK VISUAL NULL INDICATOR
INDICATING METER
FREQUENCY DIAL
PULL-TURN-PUSH ATTENUATES INPUT SIGNAL
PEAK SENS. CONTROL FOR VISUAL NULL INDICATOR
BAND SWITCH SELECTS R-F BAND
RF INPUT FOR CONNECTION TO ANTENNA OR PROBE
POWER FOR CONNECTION TO POWER SUPPLY
OSCILLOSCOPE RECEPTACLE
REMOTE METER FOR CONNECTION TO AMMETER
CALIBRATE I-F GAIN ADJUSTMENT
RECORDER FOR CONNECTION TO MILLIAMMETER-RECORDER
AUDIO GAIN CONTROL
PEAK CONTROL FOR SLIDE BACK VOLTAGE
TUNING FOR R-F STAGES
FUNCTION SWITCH
PHONES FOR CONNECTION TO HEADPHONES OR EQUIVALENT
B
SPARE FUSES
LINE FUSES
POWER ON INDICATOR LAMP
GROUND FOR EXTERNAL CONNECTION
POWER ENERGIZES EQUIPMENT
115V-230V SUPPLY INDICATOR TAG
RECORDER CHART DRIVE POWER SOURCE FOR MILLIAMMETER
POWER OUTPUT FOR CONNECTION TO RI-FI METER
POWER INPUT FOR CONNCETION TO 115 OR 230 VOLT POWER SOURCE

Figure 17-1.—Radio interference field intensity meter panel controls and receptables.

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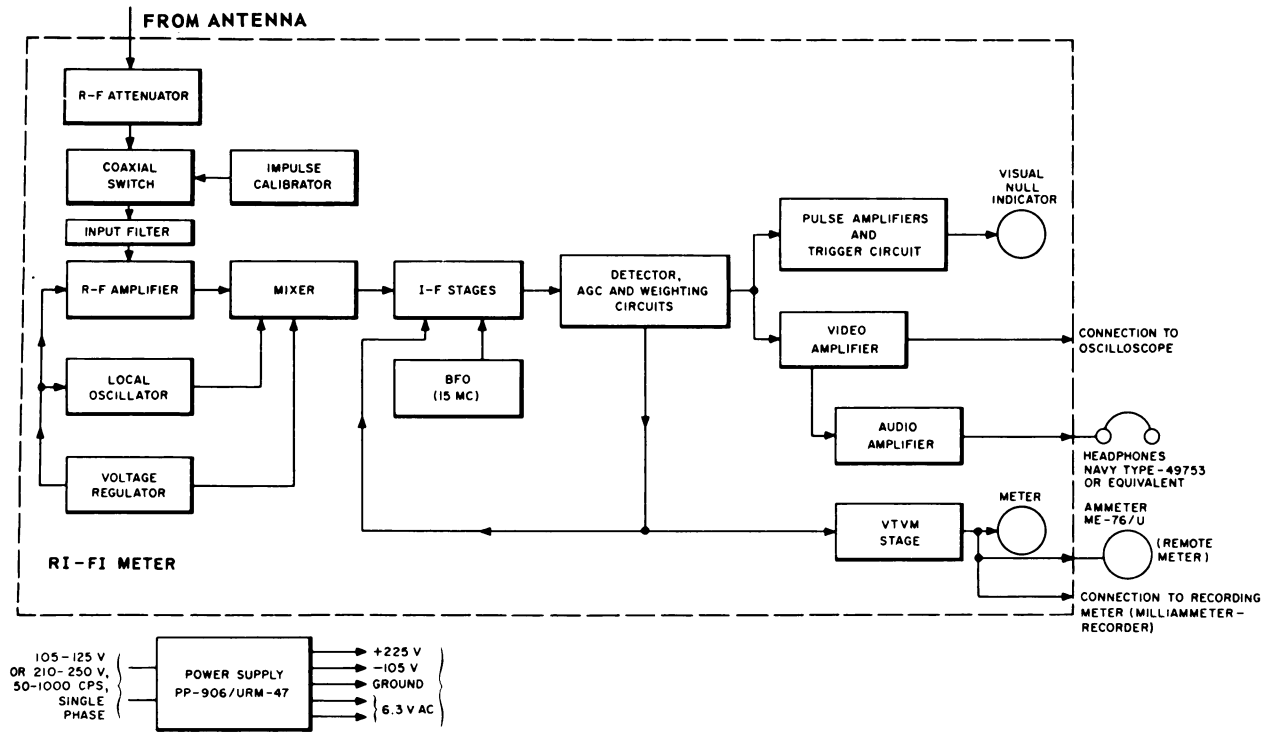

Figure 17-2.—Radio interference field intensity meter, block diagram.

The r-f signal or interferences (hereafter lled signal) picked up by the antenna may be duced in several steps through the use of a ilt-in step attenuator. The signal then passes rough a coaxial switch responsible for con- cting either the input from the attenuator or e input from a calibrator circuit to the input ter.

In order to make accurate signal strength easurements, the gain throughout the receiver ust be known. The internal calibrator supplies impulsive calibrating voltage, which may be uted through the coaxial switch into the input ter and through the remaining circuits of the ceiver. This circuit is discussed in detail ter. When the calibrator is on (function d attenuator switches on the meter front nel in the CAL position) the gain of the i-f ction can be adjusted.

When the function and attenuator switches e in any position other than the CAL position, e r-f signal from the r-f attenuator passes rough a low-pass filter, which attenuates all equencies above the tuning range of the RI-FI eter.

All input signals are amplified in the r-f stage, and mixed with the local oscillator frequency in the mixer stage. The mixer provides the intermediate frequency of 15 mc, which is amplified in five stages of i-f amplification, and demodulated in the detector stage.

The BFO crystal frequency is the same as the intermediate frequency. The signal from the BFO can therefore beat with a weak signal from the i-f stages to produce a null or near null audio note.

The strength of the signal can be accurately measured by this method with the function switch in the FIELD INTENSITY or QUASI PEAK position. The demodulated signal from the detector is acted upon by the meter detector weighting (measuring) circuits and applied to the VTVM stage. The output of the VTVM stage then activates the meter. A detailed discussion of the detector and VTVM stages appear later in this discussion.

The audio component of the signal is passed from the detector to a video amplifier, which supplies an output to an audio amplifier. Another output from the video amplifier is applied to

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an oscilloscope. The output of the audio amplifier feeds the headphones.

The pulse amplifiers and trigger circuit also receive the detector output, which is amplified and applied to a visual null indicator. The visual null indicator is provided primarily for use in peak value calibration where pulse signals are of interest. The circuit contains controls that prevent visual indication from occurring unless a positive pulse arrives at the RI-FI meter input. When properly adjusted, a peak indicator lamp flashes for each pulse input.

F-m signals may be received with reasonable audio quality through the use of "slope detection". This method simplifies circuitry, providing fair quality without the use of additional circuits and adjustments.

TYPES OF RADIO INTERFERENCE

Radio interference is defined as an electrical disturbance that causes undesirable response or malfunctioning of electronic equipment. Interference includes many forms of disturbances and may be roughly classified as continuous and discontinuous.

Continuous disturbances include random interference, such as thermal agitation. Atmospheric interference, or static, originating in lightning discharges in local storms and in more distant tropical storms, resembles random interference in that the impulses are frequent and overlap, with sharp peaks exceeding the average level.

Discontinuous interference is made up of sharp pulses, the frequency determining the character of the interference. If the pulses are relatively infrequent and clearly separated, the interference is termed impulsive. Impulsive interference may be generated by an internal combustion engine ignition system, power line discharges, electrical switching operations, electronic equipment, and by other electrical and electromechanical devices.

Because of the wide frequency range covered by continuous and discontinuous interference, the magnitude of the interference will depend upon the bandwidth of the measuring equipment. The bandwidth characteristics of the AN/URM-47 are taken into consideration and are utilized in obtaining average, quasi-peak, and peak values of interference. Shipboard electronic interference is described in detail in Volume 2 of this training course.

FIELD INTENSITY MEASUREMENTS

Field intensity is the value of the electric field at a given point and is measured in terms of volts-per-meter. One volt-per-meter is equivalent to a potential of one volt induced in a vertical rod antenna having an electrical length of one meter.

There are two principal methods of making field intensity measurements. One method is the substitution method, in which a loop antenna is connected to a receiver, and metering circuits give an indication of the receiver input. The receiver is tuned to the desired signal. The loop is rotated to give maximum signal, and the indicated value of the received signal is recorded. The loop is then rotated to give minimum received signal, and a signal generator is inserted in series with the loop. The signal generator is tuned to the received frequency and its output adjusted to obtain the same reading of signal input as was obtained with maximum loop signal. The field intensity in microvolts-per-meter is computed from the formula

$$E = \frac{V}{H(\text{eff})},$$

where V is the value of the substitute signal in microvolts and H(eff) is the effective height in meters.

The second method of field intensity measurements employed in the AN/URM-47 uses a specially designed sensitive receiver with built-in attenuators and a calibrated voltage source. The RI-FI meter is tuned to the desired frequency, and the calibrating voltage is applied to the r-f input. The calibrating signal output is standardized by adjustment of the gain in the i-f section. The calibrating voltage is then turned off, and the antenna is connected to the RI-FI meter input.

The attenuators (discussed later) are adjusted, if possible, for a meter indication at the upper half of the indicating meter dial, and the meter reading thus obtained, multiplied by the attenuator setting, (and correction factors discussed later) is the field intensity in microvolts-per-meter of the received signal. This method is advantageous in that the accuracy is independent of signal frequency, and depends primarily upon the attenuator and the linearity of the RI-FI meter circuits.

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DESCRIPTION AND USE OF ANTENNAS

igh Frequency Antenna

Either a high or low-frequency dipole, a loop robe or a ground plane vertical antenna may be sed with the Radio Interference Measuring et, AN/URM-47A. The high frequency dipole ig. 17-3) is used with frequency bands 4, 5, nd 6 (88 to 400 mc).

A coaxial line is an unbalanced transmission ne. When connecting this type line to a balanced oad, such as to the center of a half-wave ntenna, some type of balance-to-unbalance balun) converter must be used. Both the high- nd low-frequency antennas have self-contained aluns to transform the symmetrical dipole utput into an asymmetrical output suitable for oading with a 50-ohm coaxial transmission ne. A balun is actually a quarter-wave matching ection that raises the voltage of the grounded rm of the dipole (center fed) to be equal to hat of the ungrounded element. A frequency alibrated measuring tape (provided with the easuring set) may be attached to a hook at e top of the antenna so that the length of the elescoping section and balun can be measured nd adjusted to correspond to the operating equency. The balun shorting bar should be djusted so that the upper edge of the bar oincides with the frequency index on the fre- uency tape.

Eleven charts containing correction factors nd simplified operating instructions are com- onent parts of the measuring set. The charts ake up a set of calibration curves, which must e used for correcting the output indication of e equipment. A separate set of curves ac- ompanies each RI-FI, meter which are peculiar that unit alone.

The high-frequency, half-wave dipole antenna rrection factors provided on chart 4 (fig. -4) are based on an antenna in a horizontal sition and, at a height of 8-1/2 ft above the rth. When using the chart, enter from the ttom at a point corresponding to the frequency ing measured. Move upward to intersect the rve, and to the left to find the dipole cor- ction factor to be used. This factor is used in mputing signal strength measurements dis- ssed later.

Two colored telescoping extensions (fig. '-5) are contained in the high-frequency an- nna. For frequencies from 88 to approximately 6 mc, both the yellow and white extensions are used. At frequencies above 136 mc, only the use of the white extension is required (See chart 3, figure 17-5). This range, in practice, may be found to vary slightly so that an over- lapping exists at frequencies between 136 and 140 mc.

Low-Frequency Antenna

The low-frequency dipole (fig. 17-6) is used with frequency bands, 1, 2, and 3 (20 to 88 mc). Again the measuring tape is hooked to the top of the antenna so that the length of the telescoping section can be adjusted to correspond to the operating frequency. The low-frequency, half-wave dipole correction factors provided in the chart (chart No. 4, fig. 17-4) are based on an antenna in a horizontal position at a height of 10 ft above the earth.

Six telescopic extensions are contained in the low-frequency dipole. The extension required at the various frequencies in the low-frequency range are indicated in the diagram (fig. 17-5).

Fixed Length Dipole

The low-frequency dipole (fig. 17-6) is constructed so that it can also be used for field intensity measurements from 20 to 50 mc at a fixed length. This serves a particular advantage in confined places.

With the end of the measuring tape hooked at the top of the antenna, measure the telescopic length of the antenna to correspond to 50 mc (two green and two blue extensions (see chart 3, figure 17-5)). If the antenna is then to be used at 27 mc, rotate the tuning (inductance) knob on the dipole head for maximum reading on the output meter. When it is impossible to tune the knob and observe the meter at the same time, use the approximate settings for the frequency as shown in chart 3. The fixed length dipole correction factors provided in chart 5, (fig. 17-7) are based on an antenna in the horizontal position and at a height of 10 ft above the ground.

The dipole antenna must also be adjusted to a position corresponding to the polarization of the received signal. Maximum pickup of horizontally polarized waves is obtained with the dipole in the horizontal plane. Maximum pickup of vertically polarized waves is obtained with the dipole in the vertical plane. Waves of mixed polarization can be received with the dipole tilted at some angle between the horizontal and vertical planes. A set knob for locking the

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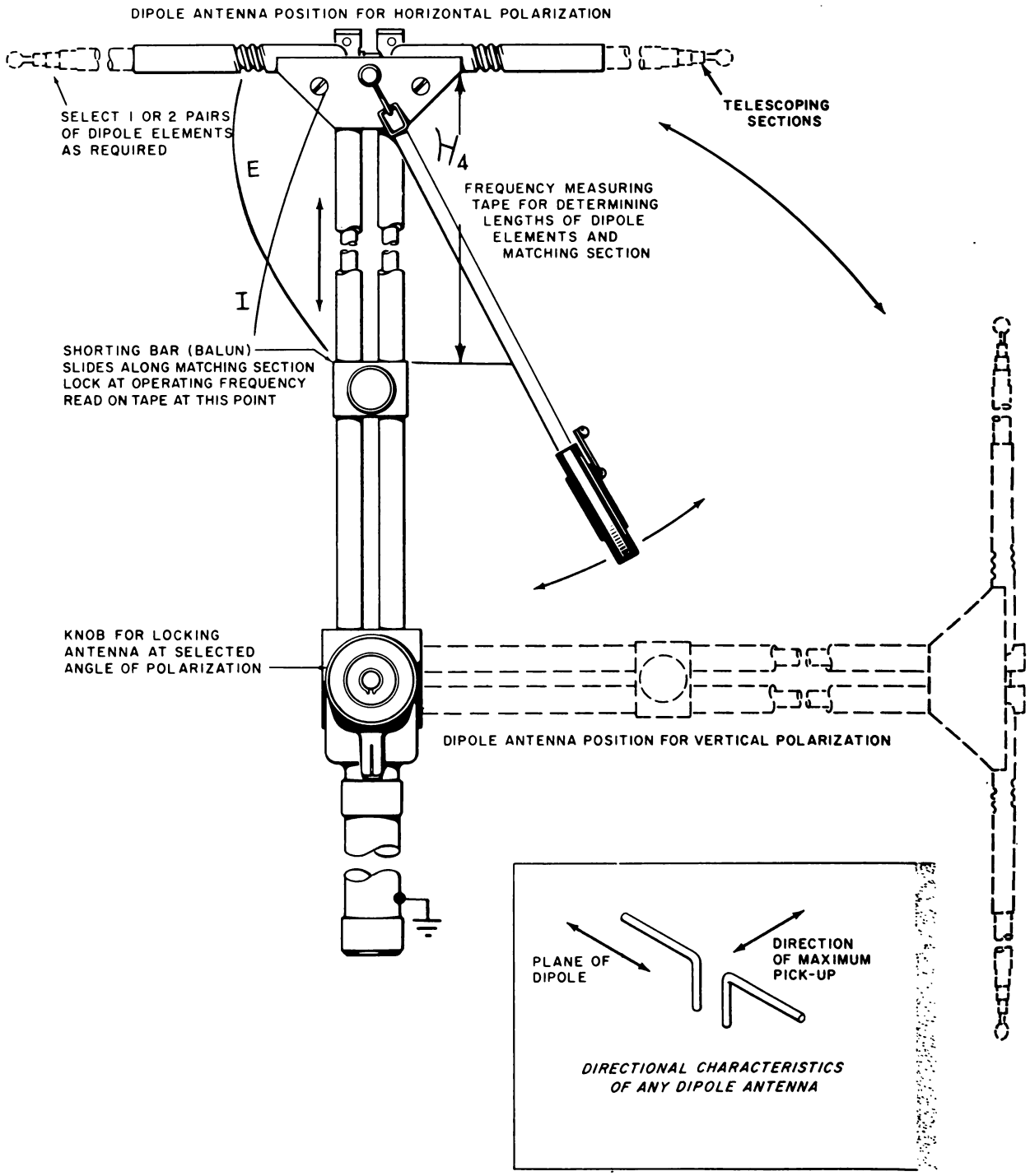
DIPOLE ANTENNA POSITION FOR HORIZONTAL POLARIZATION
SELECT 1 OR 2 PAIRS OF DIPOLE ELEMENTS AS REQUIRED
TELESCOPING SECTIONS
E
H4
FREQUENCY MEASURING TAPE FOR DETERMINING LENGTHS OF DIPOLE ELEMENTS AND MATCHING SECTION
I
SHORTING BAR (BALUN) SLIDES ALONG MATCHING SECTION LOCK AT OPERATING FREQUENCY READ ON TAPE AT THIS POINT
KNOB FOR LOCKING ANTENNA AT SELECTED ANGLE OF POLARIZATION
DIPOLE ANTENNA POSITION FOR VERTICAL POLARIZATION
PLANE OF DIPOLE
DIRECTION OF MAXIMUM PICK-UP
DIRECTIONAL CHARACTERISTICS OF ANY DIPOLE ANTENNA

Figure 17-3.—High-frequency dipole.

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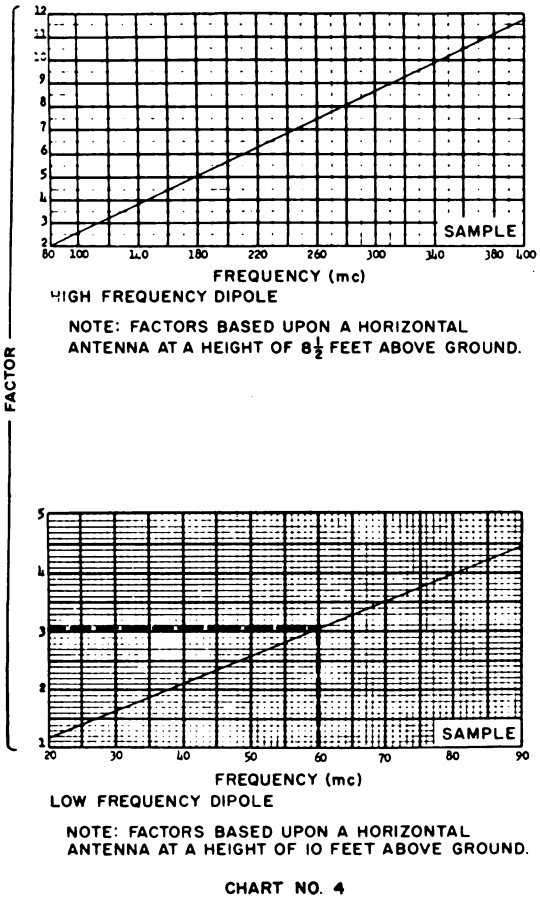
FACTOR
SAMPLE
FREQUENCY (mc)
HIGH FREQUENCY DIPOLE
NOTE: FACTORS BASED UPON A HORIZONTAL ANTENNA AT A HEIGHT OF 8½ FEET ABOVE GROUND.
SAMPLE
FREQUENCY (mc)
LOW FREQUENCY DIPOLE
NOTE: FACTORS BASED UPON A HORIZONTAL ANTENNA AT A HEIGHT OF 10 FEET ABOVE GROUND.
CHART NO. 4

Figure 17-4.—Correction factors, chart No. 4.

tenna at the selected angle of polarization is own in figure 17-6.

The directional characteristics of a half-ave dipole antenna are such that a maximum ckup exists when the dipole is broadside to e signal source. Minimum pickup is obtained nen the signal source is in line with the plane the dipole.

rtical Antenna

The ground plane vertical antenna (fig. 17-8) an omnidirectional antenna used throughout e entire frequency range of the equipment. It presents a self-resonant quarter-wave antenna the frequency range from 20 to 400 mc.

The end of the measuring tape is attached to nook at the base of the antenna, and the length the telescoping section is adjusted so that e antenna tip corresponds to the operating frequency mark on the tape. To minimize error because of variations in soil conductivity, provision is made for connecting a ground plane to the base. The function of a ground plane corresponds to that of a counterpoise used with low-frequency vertical antennas. The vertical antenna correction factors provided in chart 5 (fig. 17-7) are based on the use of 24 copper wires 160 in. long attached to the antenna base, and laid upon the ground, radiating evenly about the center. These wires are necessary when the antenna base is used on dry soil.

Loop Antenna

The loop antenna (fig. 17-9) is of small diameter (3 inches) because of the high frequencies used. The loop is used primarily for determining the bearing of a signal source, and for measuring the strength of the received signal. The directional characteristics of the loop are illustrated in figure 17-9.

The loop antenna is insulated so that it may be used as a hand-held probe for close proximity searches for sources of radiated energy. It may also be mounted on the antenna masts, using the loop base supplied with the equipment. The correction factors for the loop antenna are shown in chart 5, (fig. 17-7).

An antenna case (fig. 17-10) contains the antennas already described and all the necessary equipment for mounting and adjusting the antenna as required. Two mast sections and a tripod are provided for mounting. The two mast sections may be joined to provide increased height when necessary. The tripod has telescoping legs, which are fitted at one end with steel spikes for soft surfaces, and, at the other end with rubber feet for hard surfaces. Each leg contains two locking knobs, one for locking the legs at the desired angle, and the other for locking the legs at the desired extension.

The use of test equipment considered is of primary concern in the scope of this discussion, and throughout this chapter. In general, therefore, the discussion of circuit operation of the RI-FI meter will be limited to those circuits containing the operating controls.

CONTROL CIRCUITRY

R-F Attenuator

Signals from the antenna or pickup device are applied to the r-f input receptacle, J1 (fig.

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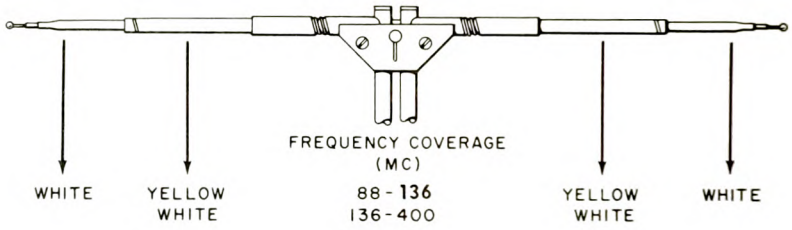
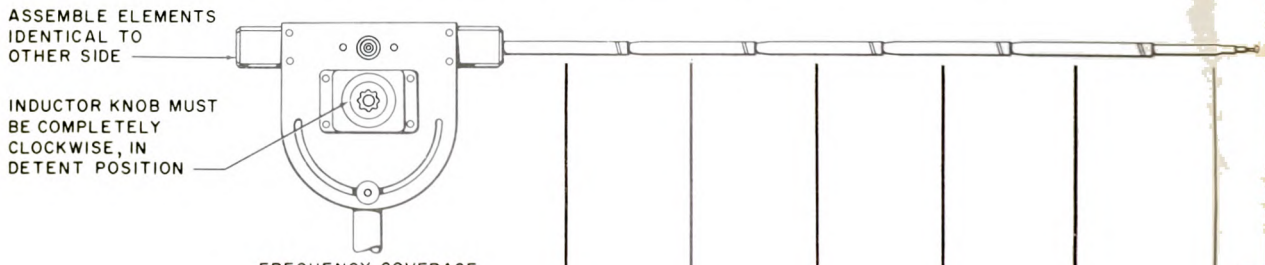
A. **HIGH FREQUENCY DIPOLE:** MAST AB-371/U (88-400 MC)



B. **LOW FREQUENCY DIPOLE:** ANTENNA AT-477/U (20-88 MC)



FREQUENCY COVERAGE (MC)						
20-23	RED	RED	RED	YELLOW	GREEN	BLUE
22-26	RED	RED	RED	GREEN	BLUE	
26-33	RED	RED	GREEN	BLUE		
32-44	RED	GREEN	BLUE			
42-63	GREEN	BLUE				
54-70	RED	YELLOW	WHITE			
70-88	RED	WHITE				

C. **FIXED LENGTH DIPOLE:** ANTENNA AT-477/U (20-50 MC)

ADJUST DIPOLE LENGTH TO 50 MEGACYCLES (USE TABLE B ABOVE). TUNE INDUCTOR FOR MAXIMUM READING OF OUTPUT METER AT FREQUENCY OF SIGNAL BEING MEASURED. ALTERNATIVELY, USE FOLLOWING APPROXIMATE VALUES OF SCALE SETTING VS. FREQUENCY.

FREQUENCY (MC)	INDUCTOR SCALE	FREQUENCY (MC)	INDUCTOR SCALE
20	2.6	35	8.1
22	4.4	40	9.0
25	5.3	45	9.5
30	7.0	50	DETENT

D. **VERTICAL ANTENNA:** ANTENNA BASE AB-364/U (20-400 MC)

THE VERTICAL ANTENNA IS CONSIDERED AS ONE-HALF OF A DIPOLE. USE DATA SHOWN ABOVE FOR ADJUSTING ONE-HALF DIPOLE. VERTICAL ANTENNA FACTORS SHOWN ON CHART NO. 5 ARE BASED ON THE USE OF A GROUND PLANE ATTACHED TO THE ANTENNA BASE CONSISTING OF 24 COPPER WIRES 160 INCHES LONG RADIATING EVENLY ABOUT THE CENTER.

CHART NO. 3

Figure 17-5.—Antenna adjustment data.

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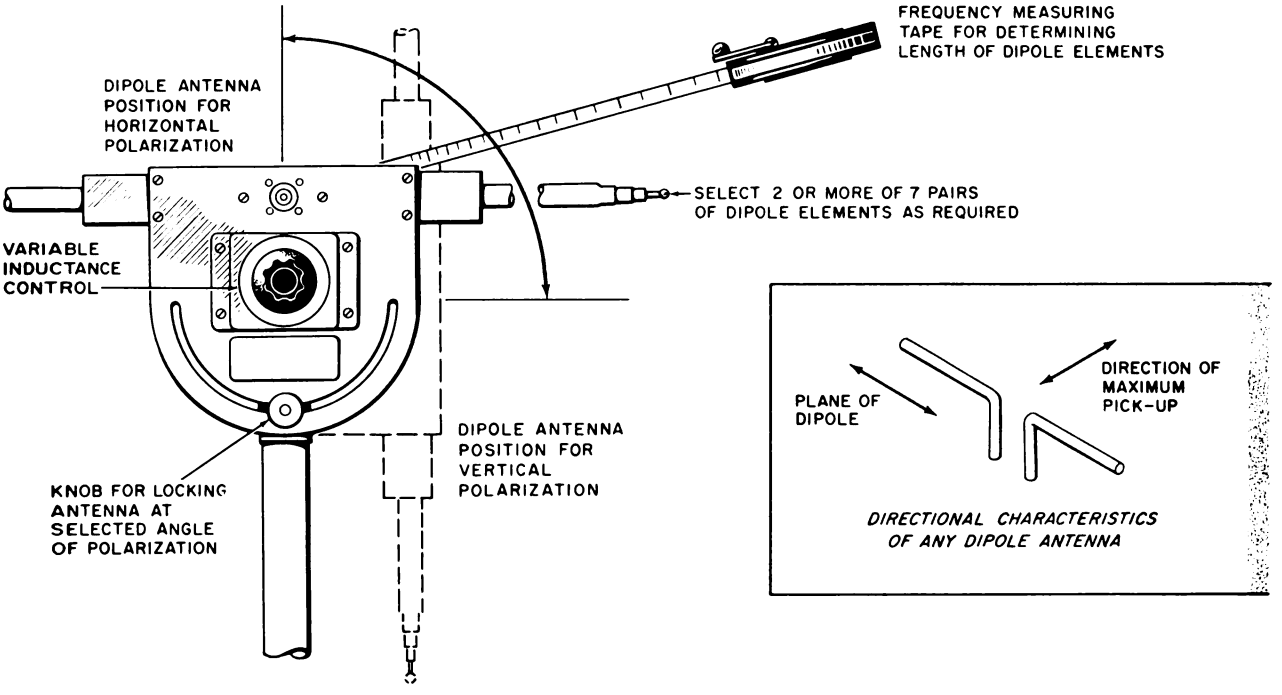
FREQUENCY MEASURING TAPE FOR DETERMINING LENGTH OF DIPOLE ELEMENTS
DIPOLE ANTENNA POSITION FOR HORIZONTAL POLARIZATION
SELECT 2 OR MORE OF 7 PAIRS OF DIPOLE ELEMENTS AS REQUIRED
VARIABLE INDUCTANCE CONTROL
DIPOLE ANTENNA POSITION FOR VERTICAL POLARIZATION
KNOB FOR LOCKING ANTENNA AT SELECTED ANGLE OF POLARIZATION
PLANE OF DIPOLE
DIRECTION OF MAXIMUM PICK-UP
DIRECTIONAL CHARACTERISTICS OF ANY DIPOLE ANTENNA

Figure 17-6.—Antenna adjustment of low-frequency dipole.

17-11). The input is then passed through a transmission line, W1, which connects the r-f input to the attenuator input jack, J2.

The r-f attenuator, Z2, consists of an assembly of six metal tubes (E1 through E6) mounted in a turret arrangement on a central shaft. The shaft, when activated, gives a rotational force to the assembly. Of the six tubes, three contain T-section resistive networks; two are "straight through" sections for zero attenuation, and one provides for the injection of a calibrating signal generated within the unit. Selection of each step of attenuation is accomplished by a pull-turn-push sequence in the activation of a shaft by means of the pull-turn-push control of the r-f attenuator.

The resistive elements sealed within the tubes are special one-watt resistors consisting of thin metallic films on ceramic discs. Their precision resistance values are maintained throughout the 20 to 400-mc range of the equipment.

The attenuator pads may easily handle one-watt of sine-wave power. However, different condition exists when pulse-type inputs are received. The difference arises because it is the peak value of the sharply pulsed signal, rather than the average power, which will damage the attenuator elements. The metallic film resistors used in the attenuator are such that with a pulse rate of 1000 pps and 4 ms duration the maximum peak voltage that can be accommodated by the attenuator is 100 v.

The six positions of the attenuator, with their corresponding amounts of attenuation are given in table 17-2. The attenuation shown is that provided in the r-f attenuator only. Additional attenuation may be provided in the i-f stages through a process discussed later.

Table 17-2.—Positions of the R-F Attenuator and Corresponding Attenuation.

Position	R-f attenuation (db) by the r-f attenuator alone
CAL	None
X1 (0DB)	0
X10 (+20DB)	0
$X10^2$ (+40DB)	20
$X10^3$ (+60DB)	40
$X10^4$ (+80DB)	60

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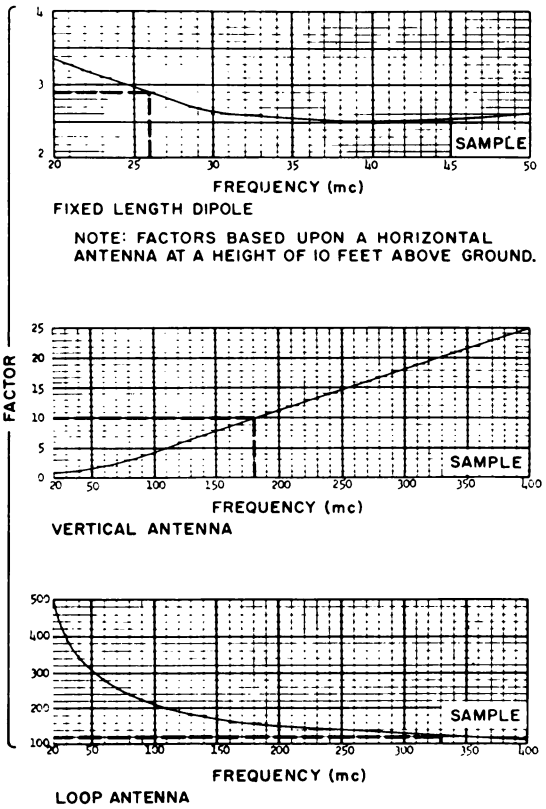

Figure 17-7.—Correction factors.

In the CAL position of the attenuator (fig. 17-11), the input signal is terminated in a 50-ohm resistive load. The terminating resistance is located within the tube, E1. By terminating the antenna signal in its characteristic impedance, leakage (which might pass through the tube to the r-f input circuits during calibration) is eliminated.

The input signal at J2 passes through one of the attenuator tubes, E2 through E6, as selected by the r-f attenuator pull-turn-push control. The signal is then passed through a coaxial switch, S2, in all positions of the attenuator except the CAL position. The S2 output is applied from J3 to transmission line W2 and to the input filter, Z4, via J4.

In the CAL position of the attenuator, a pin on the attenuator turret assembly actuates coaxial switch S2 to connect J3 to J5. This permits the calibrating signal from the impulse calibrator, Z3 (discussed later) to be injected into the r-f stage for calibrating purposes.

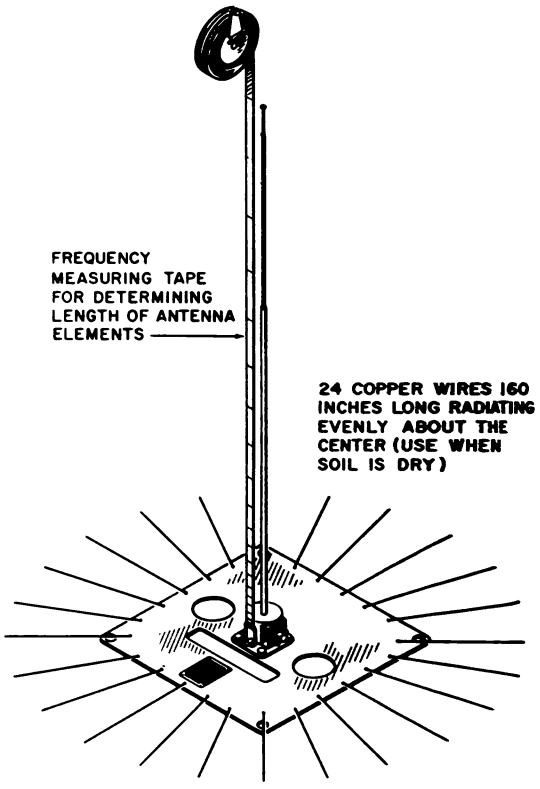
FREQUENCY MEASURING TAPE FOR DETERMINING LENGTH OF ANTENNA ELEMENTS
24 COPPER WIRES 160 INCHES LONG RADIATING EVENLY ABOUT THE CENTER (USE WHEN SOIL IS DRY)

Figure 17-8.—Ground plane vertical antenna.

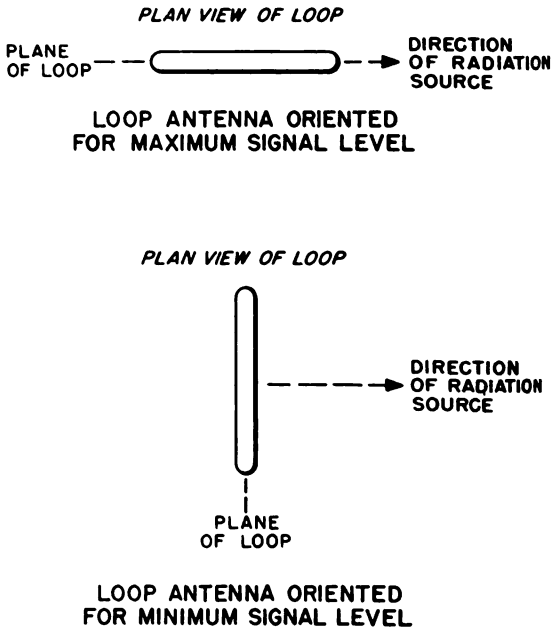
PLAN VIEW OF LOOP
PLANE OF LOOP
DIRECTION OF RADIATION SOURCE
LOOP ANTENNA ORIENTED FOR MAXIMUM SIGNAL LEVEL
PLAN VIEW OF LOOP
DIRECTION OF RADIATION SOURCE
PLANE OF LOOP
LOOP ANTENNA ORIENTED FOR MINIMUM SIGNAL LEVEL

Figure 17-9.—Directional characteristics of the loop antenna.

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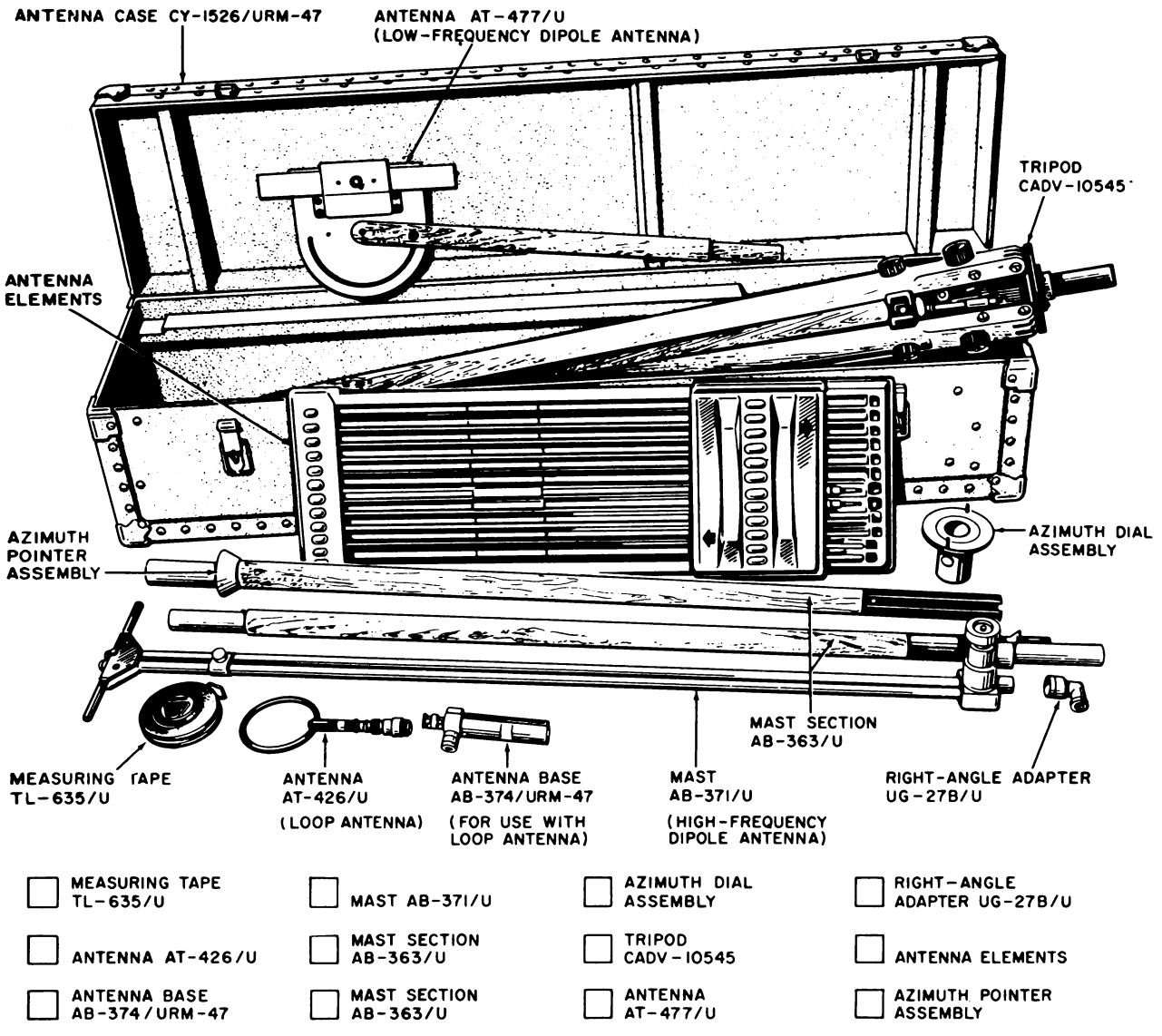

Figure 17-10.—Components contained in the antenna case.

In the X1 and CAL position of the r-f attenuator (as shown) another pin on the attenuator turret assembly activates switch S3, which opens the normally closed contacts of the switch. When S3 opens, relay K1 in the input circuit to the first i-f amplifier (fig. 17-12) deenergizes. With K1 deenergized, its contacts remove the attenuation in the i-f section as discussed later. In all positions (except the X1 and CAL position) of the r-f attenuator, S3 is closed.

The input filter, Z4, is a transmission line, low-pass filter for attenuating frequencies above the tuning range of the RI-FI meter. The inductance value of the line at these frequencies is such that it presents a large series impedance with respect to 50 ohms, while the shunt capacitance value with respect to 50 ohms is small. The line consists of six constant (below cutoff) impedance sections, including matching input and output sections. The cutoff frequency of the filter is 470 mc.

R-F Stages

The r-f stages consist of the impedance matching network, Z5, the push-pull, r-f amplifiers, V1 and V2, the local oscillator, V5, and the push-pull mixers, V3 and V4.

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A discussion of the circuit operation of these stages is not considered in this chapter. However, close observation will show that these circuits are the same as those in conventional receivers, except for the push-pull arrangement of the r-f and mixer stages.

The selected input circuit of the r-f amplifier (mounted in a turret assembly of six transformers and associated trimmer capacitors) is tuned by tuning capacitors C7, C8, C17 and C24. The tuning capacitors of all the r-f stages are ganged together and operated by the tuning control on the front panel of the RI-FI meter (fig. 17-1A). Likewise, the turret assemblies of the r-f stages are ganged and operated by the bandswitch control.

I-F Amplifier Assembly

The i-f amplifier consists of five i-f amplifiers, V8 through V12 (fig. 17-12) a second detector, V13, and a BFO, V14. The AGC and weighting circuits, the function switch, and a calibration control are also included in these stages.

The i-f signal from the r-f stages is introduced at J8 and is applied to the primary winding of the transformer, T6. The secondary of T6 is tuned to the intermediate frequency (15 mc) by the series-connected capacitors, C27 and C28.

In the X1 position of the r-f attenuator (K1 deenergized), the output from the secondary of T6 is delivered to the control grid of the first i-f amplifier, V8, through contacts 2, 4, 8, and 6 of relay K1. In this position of the attenuator, K1 is in its deenergized position (as shown). Therefore, with the r-f attenuator in this position (fig. 17-11), no attenuation is introduced to either the r-f or i-f signal. In all positions of the r-f attenuator, except X1 and CAL, relay K1 is in its energized position, and the grid of the first i-f amplifier is fed from the junction of C17 and C28 via contacts 5 and 6 of K1. Capacitors C27 and C28 form a capacitive voltage divider in all positions of the r-f ATTENUATOR, except the X1 and CAL positions, which introduce an effective fixed attenuation of 20 db into the input signal.

Note that C29 (in all positions of the r-f attenuator except the X1 and CAL positions) is connected in parallel with the secondary of T6 through contacts 1 and 2 of K1. The addition of C29 compensates for the change of the grid-to-ground capacitance of V8 across the T6 secondary, and thereby prevents detuning. R29 is connected from the junction of C27 and C28 in order to provide a d-c return path for the grid of V8.

A calibrate control is connected in the V8 cathode circuit, which controls the gain of V8, and likewise the gain of the entire i-f section. The cathode potential is obtained from a voltage divider comprising R27 and R28. R27 connects to the + 225-volt supply. The arrow across the arm of the potentiometer, R28, indicates the direction of motion of the arm for clockwise rotation. Clockwise rotation of the arm would reduce the bias on the first i-f amplifier.

Fourth I-F Amplifier Stages

The fourth i-f stage, V11 is an equalizing or gain compensation stage having its gain controlled by the different position settings of the function switch, S1A. The respective values of the selected resistors are such that the gain of the stage can be adjusted to provide the same output level from the detector stage, V13, for quasi-peak, field-intensity, and peak operation. A further consideration of the function switch is given in the discussion of the detector stage.

Detector Stage and Function Switch

The diode detector stage, V13, demodulates the i-f signal applied to its cathodes from T11. The output of the pin 7 plate is applied to two branch circuits. One of the branches passes through r-f choke L1 to the pulse amplifier and video amplifier circuits. The second branch output from the pin 7 plate is applied to the AGC and VTVM circuits in all positions of the function switch, S1C, except the QUASI PEAK position. The other diode plate (pin 2) supplies an output to the AGC and VTVM circuits when the function switch, S1C, is in the QUASI PEAK position. The quasi peak reading indicates the apparent values of peaked signals that will be discussed later.

When the function switch is in the FIELD-INTENSITY position, only one-half of the detector diode, V13, is used. R60 connects from the plate (pin 7) to contact 7 of S1C in the FIELD-INTENSITY, QUASI-PEAK, and BFO positions directly to ground, and thus serves at the diode load resistance. R61 and C57 develop the AGC potential, which is applied from C57 to the control grids of the second and third i-f amplifiers.

In the PEAK position of the function switch, S1C, peak values of pulsed signals are to be

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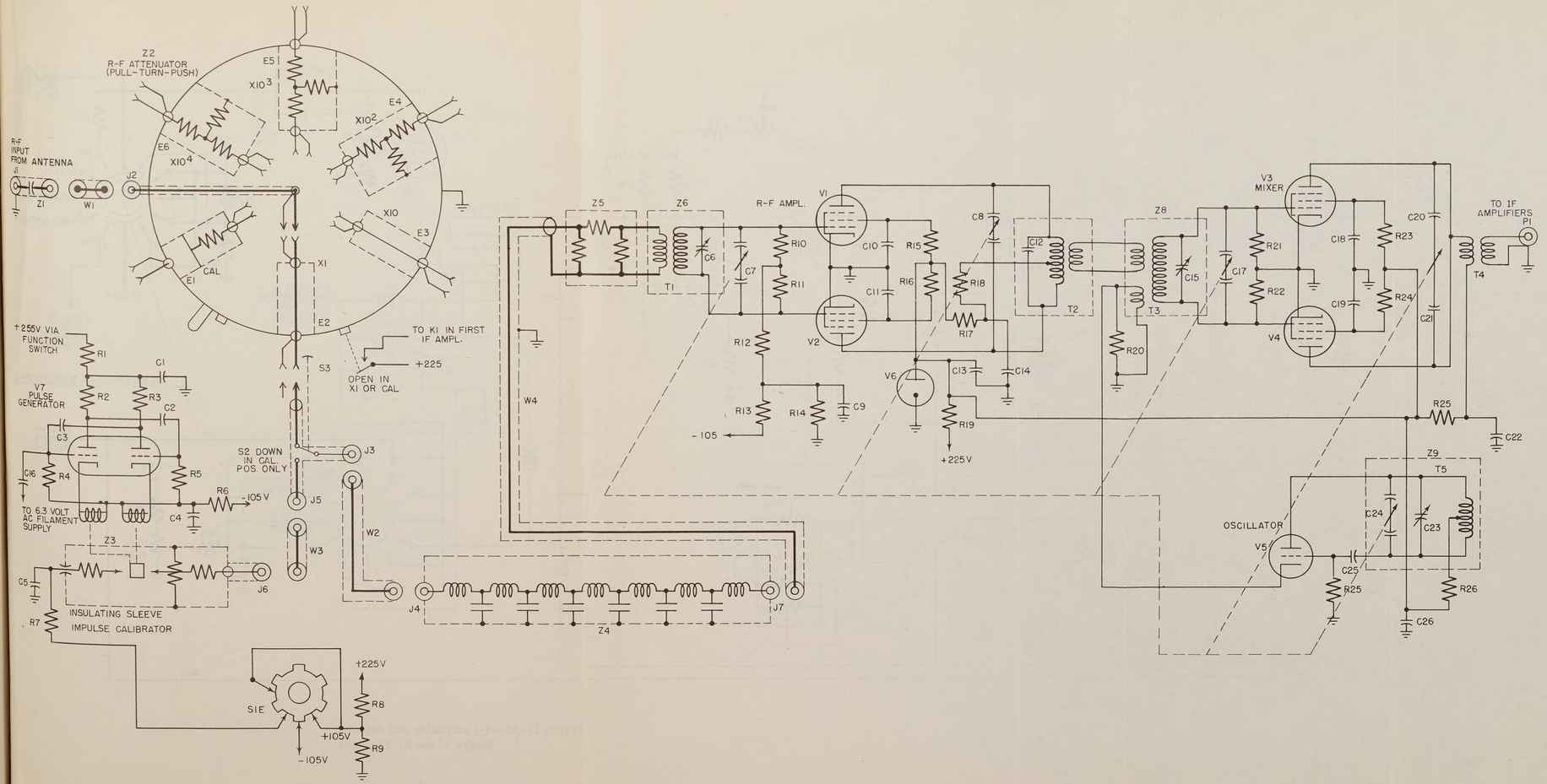

Figure 17-11.—R-f attenuator and input stages of the RI-FI meter, AN/URM-47.

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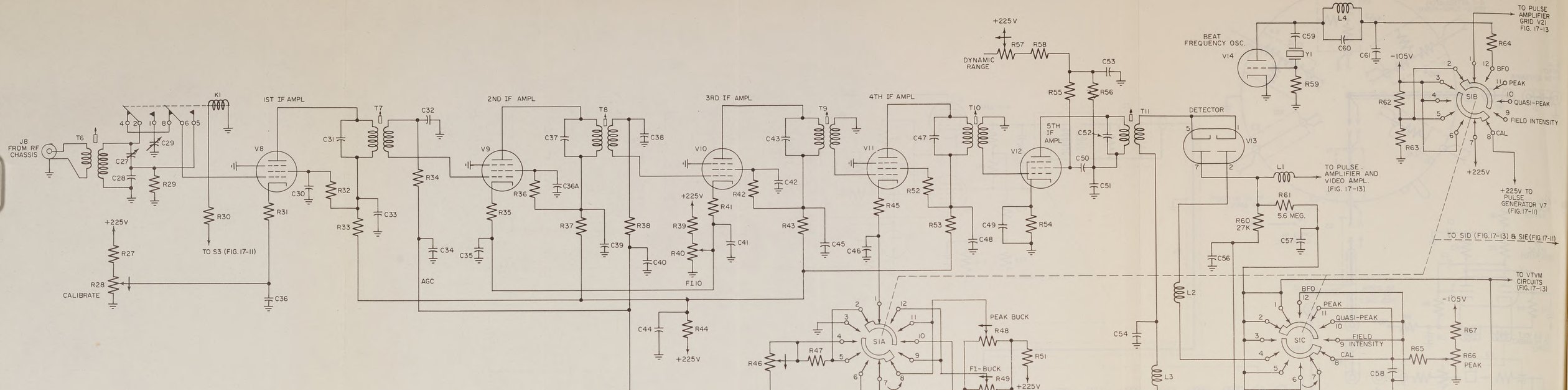

Figure 17-12.—I-f amplifier and detector stages of the RI-FI meter.

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measured. Again, only the diode section comprising pins 5 and 7 is used. R60 in this position of the switch is no longer connected to ground as in the FIELD-INTENSITY position, but is now connected through S1C (contacts 7 and 11) to a filter network composed of R65 and C58 to a negative bias control, R66. This control, called the peak control and located on the front panel of the RI-FI meter, provides the standard "slide-back" measurement facility for the determination of peak values.

In operation, the slide-back peak control, R66, is first set to produce the maximum negative bias on the detector plate (pin 7). The output of the pulse amplifier (V21 of fig. 17-13) operates the peak indicator on the front panel. The peak control R66 (fig. 17-12) is then adjusted in the reverse direction until the peak indicator just starts to flash. At this point, the signals applied to the detector are beginning to overcome the bias supplied by R66. This bias voltage value is then equal to the value of the input pulse peaks.

In the CAL position of the function switch, S1C provides the same circuit connections as in the peak position. This is done because the calibration of the equipment is accomplished by means of pulse signals from the impulse calibrator (fig. 17-11).

Function switch section S1B (fig. 17-12) provides for the application of +225 volts to the plates of the pulse generator (V7 of figure 17-11) so that the tube operates and actuates the impulse calibrator, Z3. The voltage is applied through contacts 7 and 8 of S1B.

In the BFO position of the function switch, the circuit connections are also the same as for the FIELD-INTENSITY position with the following exceptions: through counterclockwise rotation of S1D (fig. 17-13) to the BFO position, the AGC circuit is grounded through contacts 6 and 1 of this switch section. This causes the second and third i-f amplifiers to operate at maximum gain. The control grid of V19 (in the VTVM circuit of figure 17-13) is also grounded through the same connection to disable the circuit and protect the indicating meter from being driven off scale. The fourth i-f amplifier V11, (fig. 17-12) is permitted to operate at maximum gain since its cathode resistor R45, is now directly returned to ground through contacts 6 and 1 of S1A. S1B allows the application of the +225 volts to the plate of the BFO tube, V14 through contacts 7 and 12, permitting the tube to operate.

Both sections of the detector diode, V13, operate when the function switch is placed in the QUASI-PEAK position. The term, quasi, is defined as a certain sense of degree, or a seeming value. As used in this discussion, quasi peak refers to the seeming or apparent value of a chain of pulses that appear in rapid succession.

R61 is added in series with R60 (the diode load resistor) in the QUASI-PEAK position of the function switch through the conduction of the second diode (pins 1 and 2). The conduction path for this diode is through L2, contacts 4 and 1 of S1C, through R61 and R60, and contacts 7 and 10 of S1C to ground. Resistors R61 and R60 and capacitor C57 constitute the weighting circuit. The term, weighting, refers to the time constant introduced in the detector and AGC circuits. The term may be better described as the time versus voltage characteristics of the detector output circuits.

When the detector conducts, C57 charges up rapidly from ground through the arm of R50, through S1A (contacts 10 and 7), through L3 and the secondary winding of T11, and through the conducting diode section of V13 comprising pins 1 and 2, through L2 and contacts 4 and 1 of S1C, to C57 and ground (bypassing R61 and R60). The charge time is approximately 1 msec. During intervals when V13 is not conducting, C57 discharges slowly through R61 and R60, and through contacts 7 and 10 of S1C to ground. Owing to the large ohmic value of the two series resistors, the weighting circuit employs a single clamping circuit principle with which you are already familiar. The discharge time (RC) of C57 is approximately 600 msec.

When the received signal contains pulses occurring in rapid succession, the average voltage appearing across R60 and R61 will be near the peak value of the pulses. This voltage is used as the AGC and quasi-peak value which is applied to the grids of the second and third i-f amplifiers, V9 and V10, and to the VTVM tube (V19 of figure 17-13) respectively.

When receiving unmodulated r-f input signals, the output indications on the output meter (discussed later) should be the same for both field-intensity and quasi-peak operation. Because of the addition of the weighting circuits in the QUASI PEAK position of the function switch, it becomes necessary to adjust the bias in the cathode of the fourth i-f amplifier, V11 (fig. 17-12). This is accomplished by an adjustable resistor, R46 (quasi peak 100), which is

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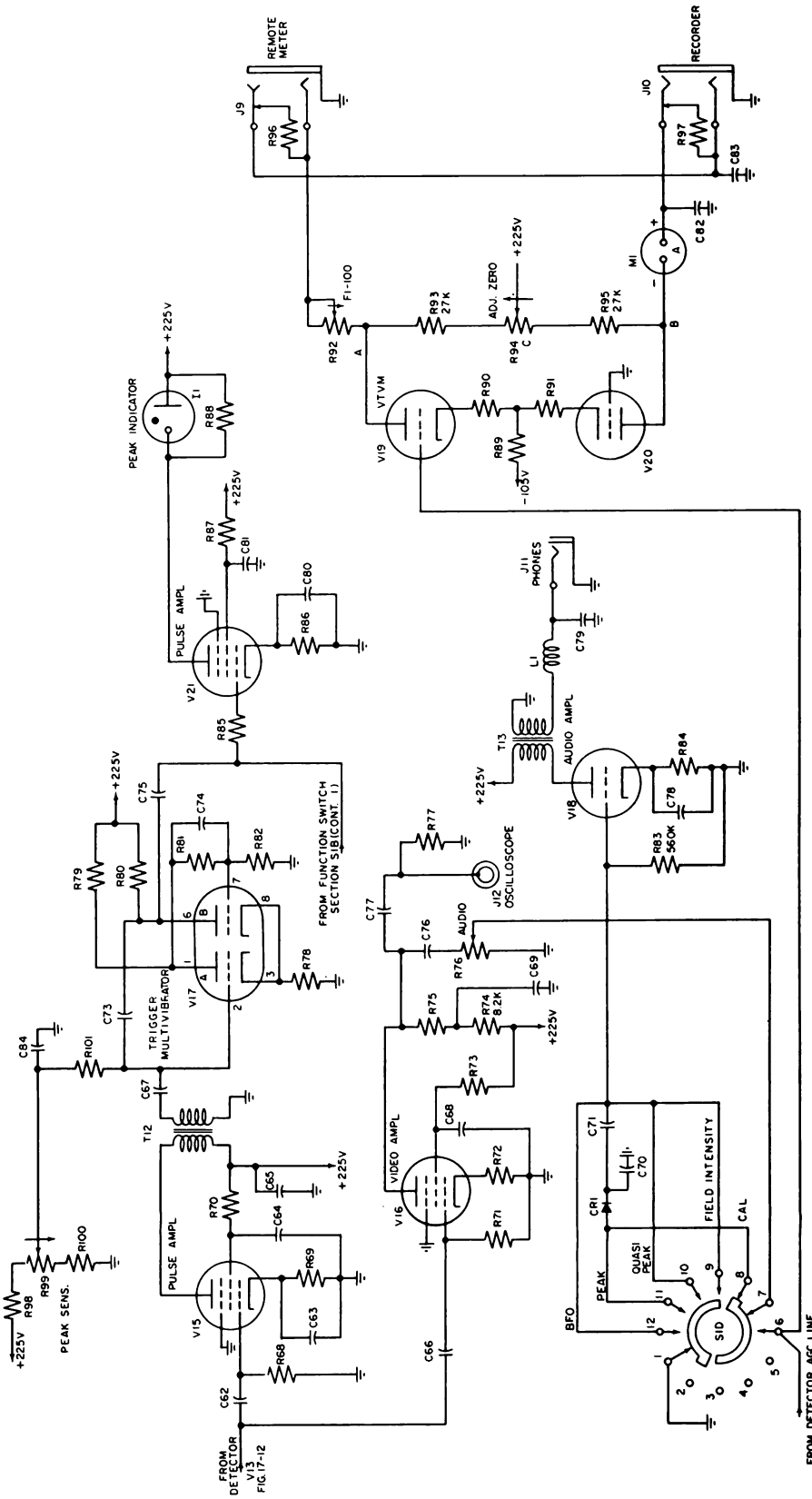

Figure 17-13.—Output stages of the RI-FI meter AN/URM-47.

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placed in the circuit through contacts 1 and 4 of S1A.

A small contact voltage will exist between the plate and cathode of V13 when no signal is introduced. The effect is one of varying initial electron velocity in the diode caused by thermionic emission from the cathode. Some of the electrons pass to the plate and are evidenced by a voltage drop across the diode plate load resistor. If the condition was not corrected or counteracted, this voltage (the diode self-bias potential) would be passed from the load resistor to the VTVM and indicating meter circuit, causing a certain amount of error deflection in the output reading.

A small positive voltage, derived from the FI-BUCK control, R49, is applied through the function switch section S1A in the FIELD-INTENSITY position (contacts 9 and 7), through L3 and the T11 secondary to the cathode of V13. The FI-BUCK control, R49, is adjusted so that it just cancels the self-bias potential of V13 and the meter reading on the indicating meter drops to zero.

The efficiency of the detector diode in quasi-peak operation differs from that in field intensity because of the difference in the value of load resistance. Likewise, a different value of self-bias potential will exist between the cathode and plate of V13. To counteract this effect, the cathode of V13 receives a positive potential from the quasi-peak 1 to control, R50.

Function switch, section S1A, in the CAL and PEAK positions permits application of a positive bias from potentiometer R48 (the peak buck control) to the detector, V13. This is done to equalize the indicating meter reading for low-signal levels in the PEAK position with that obtained in the QUASI PEAK position of the function switch. S1A in the PEAK and CAL positions also introduces an additional resistor, R47, in the cathode circuit of the fourth i-f amplifier, V11, to modify the gain of this stage in accordance with peak operation.

PULSE AMPLIFIER AND VISUAL NULL (PEAK) INDICATOR STAGES

The pulse amplifier, V15 (fig. 17-13) receives its input signal from the detector stage, V13 of figure 17-12. The input signals are either pulses or sine waves, which, after amplification, are used for oscilloscopic and aural indication, respectively.

The visual null (peak) indicator, I1, is used primarily in peak value calibration where pulse signals are of interest. Accordingly, the pulse amplifier, V15, and its associated circuits provide greater gain for pulse signals than for sine waves.

The negative signals from the detector (V13 of figure 17-12) are developed across C62 and R68, and applied (from the junction of the two) to the control grid of the pulse amplifier. The positive (or inverted) output pulses, taken from the plate of V15 are coupled through T12 to the control grid of the multivibrator, V17.

The trigger multivibrator shapes the input pulse and provides it with the necessary duration for operating the peak indicator. Tube V17 is connected as a one-shot multivibrator trigger circuit. Signals from the pulse amplifier, V15, are inverted by the transformer, T12, and applied through C67 to the control grid (pin 2) of the trigger multivibrator, V17.

Both of the control grids of V17 (pins 2 and 7) are held positive with respect to ground through their respective connections to the B supply. Both sections of the tube, however, will not conduct at the same time because the conduction of the A section will develop a bias voltage from cathode to ground, which is sufficient to cut off the B section. It should be noted that the grid to cathode voltage (bias) of the A section is determined by the setting of the arm of R99. The bias on the B section of V17 is determined by the potential from cathode to ground (developed across the common cathode resistor R78 and grid resistor R82) which the A section is conducting.

The circuit is so arranged that the pin 2 grid to ground voltage is approximately 36 v. The common cathode to ground voltage is about 47 v, and the pin 7 voltage to ground is approximately 18 v. Therefore, the grid bias of the A section of V17 is -11 v (allowing it to conduct) while the grid voltage (bias) of the B section is approximately -29 v — a value sufficient to hold the B section beyond cutoff.

Negative pulses coupled through C67 from the pulse amplifier will drive the A section of V17 to cutoff. The current through R78 will decrease, thereby decreasing the cathode to ground voltage. Likewise, the positive-going plate of V17A causes C74 to charge through R82, making the control grid of the B section of V17 still more positive. The simultaneous actions are sufficient to cause V17B to conduct.

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As the pulse at the pin 2 grid becomes less negative, V17A again conducts with an accompanying drop in plate voltage and increase in cathode bias. The discharge of C74 through R82 aids the increased bias to cut off V17B. As V17B cuts off, its plate potential begins to rise, charging C75 through R63 (fig. 17-21) in the CAL and PEAK positions of the function switch (section S1B). The circuit time constant is such that a positive pulse of approximately 1/20 second duration appears at the pin 6 plate for each negative input pulse from the pulse amplifier, V15. The output pulse of V17B is coupled through R85 (a grid limiting resistor) to the grid of the pulse amplifier, V21.

Peak Indicator

In the BFO, QUASI-PEAK, and FIELD-INTENSITY positions of the function switch (section S1B of figure 17-12) -105 v is applied through the sliding contact (pin 1) and through R85 (fig. 17-13) to the control grid of V21. This potential is sufficient to hold V21 beyond its cutoff point for these modes of operation, and the peak indicator is not activated. In the PEAK and CAL positions of S1B, the pulse amplifier, V21, is provided with a suitable bias (approximately -12 v) derived from the voltage divider (formed by resistors R62 and R63 in figure 17-12), which is connected between the -105 v source and ground. This negative bias prevents the conduction of V21 until a positive pulse arrives from the trigger multivibrator, V17.

When the positive pulse arrives from V17B, (fig. 17-13) the grid cathode voltage of V21 is reduced by the required amount to allow the tube to conduct. As V21 conducts, a current passes through the peak indicator, I1, for the duration of the output pulse. The peak indicator lamp therefore flashes for each pulse output of V21.

VTVM and Indicating Meter Stage

The vacuum tube voltmeter circuit (fig. 17-13) consists of a balanced bridge circuit in which tubes V19 and V20 form two adjacent arms. The d-c voltage to be measured is applied to the grid of V19, which upsets the balance of the bridge. The input voltage is received from the detector stage, V13 (fig. 17-12).

When the bridge is properly balanced, the current flow through R93 and the upper portion of R94 is exactly equal and opposite to that through R95 and the lower portion of R94. In this condition, the net voltage existing between points A and B is zero. R92, R96, R97, and M1 are connected in series across the A and B terminals. Since the voltage applied to the network is zero, the current passing through the meter is also zero, and the meter indicates zero.

The cathodes of V19 and V20 are returned to the -105-v supply through R89 (and the respective cathode resistors) so that the tubes conduct. Negative input signals from the detector cause the current through V19 to decrease in accordance with the amplitude of the detector rectified output. This, in turn, upsets the balance in the voltage, AC and BC, so that a difference in potential exists between terminals AB. A current proportional to the unbalance passes through the meter, which indicates the amplitude of the RI-FI meter input.

The adjust zero control, R94, is used to adjust the voltage applied to the plates of V19 and V20 for balancing of the bridge. Balance is indicated by a zero indication on the meter, M1, in the absence of a signal input. The FI-100 control, R92, in series with the meter, M1, is a voltmeter sensitivity control for adjusting the meter full-scale deflection.

The remote meter jack, J9, and the recorder jack, J10, provide connection facilities for a remote indicating meter and/or recording meter in series with meter M1. (R-f pickup by cables connected to these jacks is filtered by C82 and C83.)

Pulse Generator and Impulse Calibrator Stages

The pulse generator, V7 (fig. 17-11) and impulse calibrator Z3 constitute a standard calibrating source contained within the equipment. Output signals from the calibrating circuit are coupled to the RI-FI meter input stages when the function switch is set to the CAL position, as discussed earlier.

The pulse generator is essentially a free running multivibrator operating at a low repetition rate of approximately 8 to 20 PPS. V7 is energized through contacts 7 and 8 of function switch section, S13 (fig. 17-12) when the switch is in the CAL position. C16, connected between the control grid and the filament supply of V7, is used for the application of an initial disturbance voltage to the grid to start the action of the multivibrator. The pulse generator drives the impulse calibrator by means of coils, which are connected in the V7 cathode circuits.

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The impulse calibrator is contained in a brass tube. The tube is insulated from its inner components by an insulating sleeve. An iron slug with platinum contacts at either end moves axially (parallel to the sleeve) between two mating platinum contacts, which are also contained in the insulating sleeve.

The sections of the multivibrator tube, V7, alternately conduct. The current through the cathode coils is such that one coil is active while the other is dormant. The position of the iron slug in the insulating tube is controlled by the coils in such a way that when the left coil is active, the iron slug is caused to move to the left. Conversely, when the right coil is active the iron slug moves to the right.

The movement of the slug in its left and right motion causes it to alternately mate with two other platinum contacts. When the slug is moved to its extreme left position, a charge (with respect to ground) is placed on the slug from either a +105-v or -105-v source, through the contacts of S1E. The +105 v is obtained from a voltage divider comprising R8 and R9 connected between the +225-v supply and ground. When the slug is caused to move to the extreme right position, it makes contact with a 10-ohm disc resistor through which the charge of the slug passes to ground. The discharge of the slug creates a voltage pulse at the output terminal of Z3.

The output pulses of the impulse calibrator are delivered through jack J6, transmission line W3, and jack J5 to the coaxial switch, S2. Since the r-f attenuator is set in the CAL position, the notch on the turret assembly causes S2 to move to the DOWN (or CAL) position. The output of the impulse calibrator is therefore coupled through J5 to J3 and to the input circuits of the RI-FI meter. These pulses are of constant known amplitude and constitute the calibration signal for the equipment.

To prevent the transfer of metal at the platinum contacts, which would occur if the polarity of the charging voltage were constant, the polarity is reversed each time the pulse generator and impulse calibrator are initially energized. The action is accomplished by S1E, which is ratchet driven by the function switch shaft. One position of the switch connects the wafer section to the -105-v supply. The next position of the switch connects the wafer section to the +105-volts. Subsequent positions of the switch connect the wafer section alternately to -105 v and +105 v.

RADIO INTERFERENCE CONSIDERATIONS

Selecting the Site for Interference Surveys

An open, flat terrain is preferred (when one is making field intensity measurements) to avoid the possibility of absorption by dense growths of trees or steep ridges. Also, the face of a nearby cliff may act as a reflector and produce local variations in the intensity of the interference measured.

To isolate the radiated field from the induction field of a source of interference, make radio interference measurements of a radiating source at distances greater than twice the wavelength of the radiated signal.

Interference measurements made near power lines are subject to inaccurate results because of possible reradiation from the power lines. Every attempt should be made to avoid locating the equipment near closed loops present in iron frame buildings and topside structures aboard ship. If possible, avoid locating in the vicinity of underground pipes, trolley cables, or rail lines.

When the equipment is installed aboard ship or in a vehicle, it should be remembered that the metal body and frame of the vehicle tend to distort the pickup pattern of the dipole antenna. In addition, closed loops formed by wiring or structural members will affect the field pattern. In some cases, these errors can be determined with the equipment mounted in a designated location. Correction factors so obtained may then be applied to compensate for pickup pattern distortion.

When a variety of equipment, including radio receivers, are employed at any one site, a check for interaction between equipments should be conducted.

The ideal site for making interference surveys is an open, flat terrain at a considerable distance (1000 ft or more) from buildings, electric lines, fences, and other wave obstructing devices. Ideal sites are rare in the more populated sections of the country. Therefore, it is good practice to check a proposed location by making signal measurements of the desired signal at several points in the vicinity. If the same value of field intensity is obtained at each of the points, any one may be considered satisfactory. If it is necessary to use an unsatisfactory site, a series of readings should be recorded at a number of different points in

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the neighborhood of the selected site, and detailed notes on the site conditions should be appended to the recorded data.

Making An Interference Survey

An interference survey of a suspected area should begin with a series of measurements, at the frequencies under investigation, made with the appropriate pickup. It is customary in general surveys to make measurements at five frequencies in each band of the equipment with sufficient overlap to insure complete coverage of all bands. If a station or service is found operating on the frequency selected for measurement, move sufficiently away in frequency to avoid side bands of the station. The survey report should, however, note the frequency and field intensity of signals from the station.

Certain types of interference surveys require scanning an entire band for evidence of peaks or minima. Any such peaks or minima found must be measured and noted in the survey report.

During the survey, the received signals should be aurally monitored at a volume level that will enable identification of the signal or interference.

Three-point fixes on each source of interference enable the geographic location to be determined by triangulation. Investigate each interference source, moving the RI-FI meter as close as is practical to the source to ascertain the distant fix. When the location has been narrowed down to a small area, use the loop probe to further determine the location of the offending equipment or power line.

If the interference is found to be a power line or other current-carrying conductors, use the RI-FI meter as a two-terminal voltmeter to determine the amount of conducted interference on the line. Standard safety precautions should be considered when connecting to power lines of any equipment, which may be energized from a switch not under the control of the operator of the radio interference measuring set. All connections should be made with the power disconnected from the lines of the equipment under test.

Standardizing Gain

After turning on the power switch (fig. 17-1, A):

1. Set the bandswitch (G) and the tuning dial (H) to the desired frequency. For best results, allow a 30-minute warmup period after turning the power on.

2. Pull the attenuator knob (A) out, and turn it to the CAL position. Afterwards, push the control all the way in to avoid possible discontinuity at the coaxial connectors of the attenuator.

3. Rotate the function switch (B) to the CAL position. This position is used for standardizing the receiver gain, and for placing the calibrator circuit in operation, as discussed.

4. Rotate the peak control (C) (R66 in figure 17-12) to obtain approximately full-scale meter reading. The peak control provides the slide back voltage for making peak measurements as discussed. It should be recalled that the impulse calibrator circuit provides impulsive signals that operate the peak indicator.

5. Rotate the calibrate control (D) to the maximum counterclockwise position. This action places maximum bias on the first i-f amplifier grid (V8 of figure 17-12) to insure complete blocking of input signals to the RI-FI meter during calibration.

6. Adjust the peak sensitivity control (E) clockwise until the peak indicator lamp glows. This action triggers the multivibrators (V17 of figure 17-13) which, in turn, causes V21 to pass a current through the peak indicator, I1. The peak sensitivity control is then adjusted counterclockwise until the lamp just ceases to glow. The visual indicator circuit is then set at its most sensitive position.

When making measurements on a given frequency, it is necessary that you adjust (calibrate) the gain of the RI-FI meter stages at the desired frequency. A calibration reading is taken from charts 1 and 2 (figure 17-14), which contain six curves covering the six frequency bands. The curves shown are samples, and should not be used for actual calibration of the equipment. In practice, separate charts are used for the calibration of impulsive and sinusoidal signals.

7. Obtain the correct value of calibration voltage from charts 1 and 2 for standardizing the equipment at the desired frequency. For convenience, the voltage is indicated on the chart in db. Adjust the peak control (C) until the indicating meter (fig. 17-13) reads the correct value of calibration voltage on the decibel scale.

For instance, if a signal measurement of 27 mc is to be considered, enter the chart No. 1 (fig. 17-14) from the bottom at 27 mc on the

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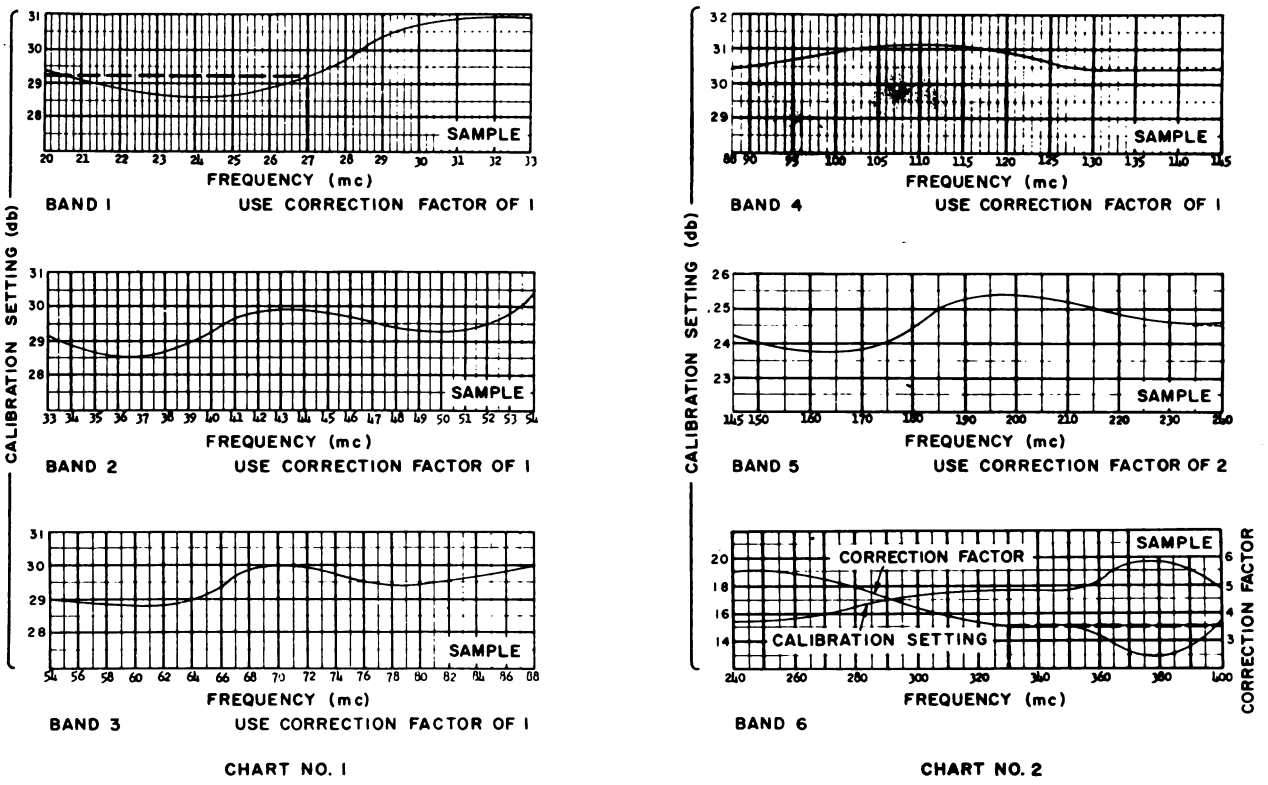

Figure 17-14.—Calibration settings (charts Nos. 1 and 2).

band 1 curve. Move upward to intersect the plotted curve. From the point of intersection, follow to the left to find the calibration setting to be used, which, in this case, is 29.4 db. Adjust the peak control until the meter reads the correct calibrating voltage on the db scale (29.4 db).

8. Adjust the calibrate control (D) clockwise (reducing the first i-f amplifier bias) until the peak indicator lamp blinks in step with the calibrator pulses. Adjust the control counterclockwise until the indicator lamp just ceases to light. The RI-FI meter gain is now standardized at this frequency. The calibrate control setting should not be changed until the RI-FI meter is again standardized for a gain at this or another frequency.

Measuring Sine-Wave Signals

Before making any type measurements, the equipment must be standardized as discussed above. After this procedure has been accomplished, the measurement of sine-wave voltages should begin by turning the function switch (B) to the FI position for a rms reading of a c-w signal, or to the QP or PEAK position when the carrier plus modulation is to be measured.

Adjust the position of the attentuator control (A) by the pull-turn-push sequence until the meter reads within the scale range, preferably in the upper portion of the meter scale.

The signal being measured should be close to the frequency of gain standardization, as described above. Otherwise, it will be necessary to standardize the gain at the new frequency before a measurement can be accurately made.

To determine the signal strength, read the indicating meter in microvolts, and multiply this reading by the correction factor obtained from chart 1 or 2. This product is then multiplied by the attenuator control setting. For example, assume the attenuator control setting to be $X10^2$, a meter reading of 20 μv and a correction factor of 1. Then, 20 x 1 x 10^2 = 2000 μv across a 50-ohm impedance.

Other Applications of the RI-FI Meter

The use of the RI-FI meter previously discussed represents a single test application of

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the equipment. With a knowledge of the function and circuits of the operating controls considered earlier, the ET should recognize the fact that the RI-FI meter makes itself applicable to many test situations. To a great extent, the number and variety of these applications are limited by the skill of the operator.

Other uses of the RI-FI meter include:

1. Measurement as a null instrument in conjunction with r-f bridges and slotted lines, which operate in the 20 to 400-mc frequency range. Extremely short leads, preferably shielded, must be used and care must be taken to assure proper impedance matching.

2. Checking the gain of a radio frequency stage in radio equipment by measuring the voltage input to the stage and the signal output of the stage. It should be noted that the RI-FI meter shunts approximately 50 ohms across the circuit and may upset the stage under test unless appropriately isolated.

3. Checking the voltage of any 50-ohm transmission line up to one volt, provided the transmission line is properly terminated. The voltage rating (d-c or a-c power frequencies) that may be applied to the input without damage to the input circuits is 1000 v.

4. Determining the loss in a coaxial line connected to a signal generator by comparing the signal level at the input and output ends of the line. The appropriate impedance matching network must be used between the line and the RI-FI meter. To minimize the effects of standing waves, the higher attenuation positions ($X10^2$, $X10^3$, and $X10^4$) should be used when practicable.

OPERATION OF A REPRESENTATIVE SYNCHROSCOPE

The synchroscope (cathode-ray type) is an oscilloscope having a fast sweep triggered by a synchronizing signal. The synchroscope considered in this discussion is that of the AN/USM-32. This unit was chosen because it presently receives wide usage by Naval electronics personnel, and, since its circuits represent, to a great extent, those common to most of the present-day synchroscopes.

The oscilloscope of the AN/USM-32 (fig. 17-15) is a general purpose maintenance instrument, which presents electrical waveforms on the face of a three-inch cathode-ray tube. Signals to be analyzed are normally connected to the vertical signal terminal from which they pass through built-in amplifiers. Through the use of these amplifiers, a signal as low as 0.28 volt peak-to-peak (0.1 volt rms) will produce an inch of deflection on the oscilloscope cathode-ray screen. The vertical direct terminal provides for direct signal connection to the cathode-ray tube deflection plates, in which case 43 to 52 peak-to-peak volts will produce one inch of deflection.

BLOCK DIAGRAM

The synchroscope of the AN/USM-32 can be considered as consisting of six major sections. These are the vertical deflection system, the horizontal deflection system, the amplitude calibrator, the time calibration circuit, the trigger generator, and the power supply. The block diagram of figure 17-16 shows how the circuits are related in the oscilloscope.

Input signals from the vertical signal terminal are first applied to the input attenuator. The position of the attenuator is controlled by the volts-per-inch control (fig. 17-15), which indicates the amount of voltage required for one inch of deflection. The signal voltage attenuator and vertical gain control permit the application of as high as 600 v d-c to the vertical signal terminal without producing off-screen deflection. The input stage is connected as a cathode follower to present a high impedance to the circuit under test.

Provision is made for delaying the signal through the vertical amplifiers. This is done to allow signals with a fast rise time to be observed on the sweep trace.

The vertical direct terminal provides for direct signal connection to the cathode-ray tube deflection plates. When this connection is used, 43 to 52 peak-to-peak volts will produce one-inch deflection.

A driven-type sweep circuit is employed in the oscilloscope, which can be triggered by the amplified input signal, an externally applied sync signal, an internally available trigger, or the line frequency signal. The choice of the driving signal is determined by the position of the synchronization switch. The sweep duration is continuously variable from 10 to 200,000 μs. The beam is brightened during the forward portion of the sweep.

Time calibration signals are provided for accurate time measurement of the visual presentation. Timing marker intervals of 1, 10, 100, 1000, and 10,000 μs may be selected by the marker interval control on the oscilloscope

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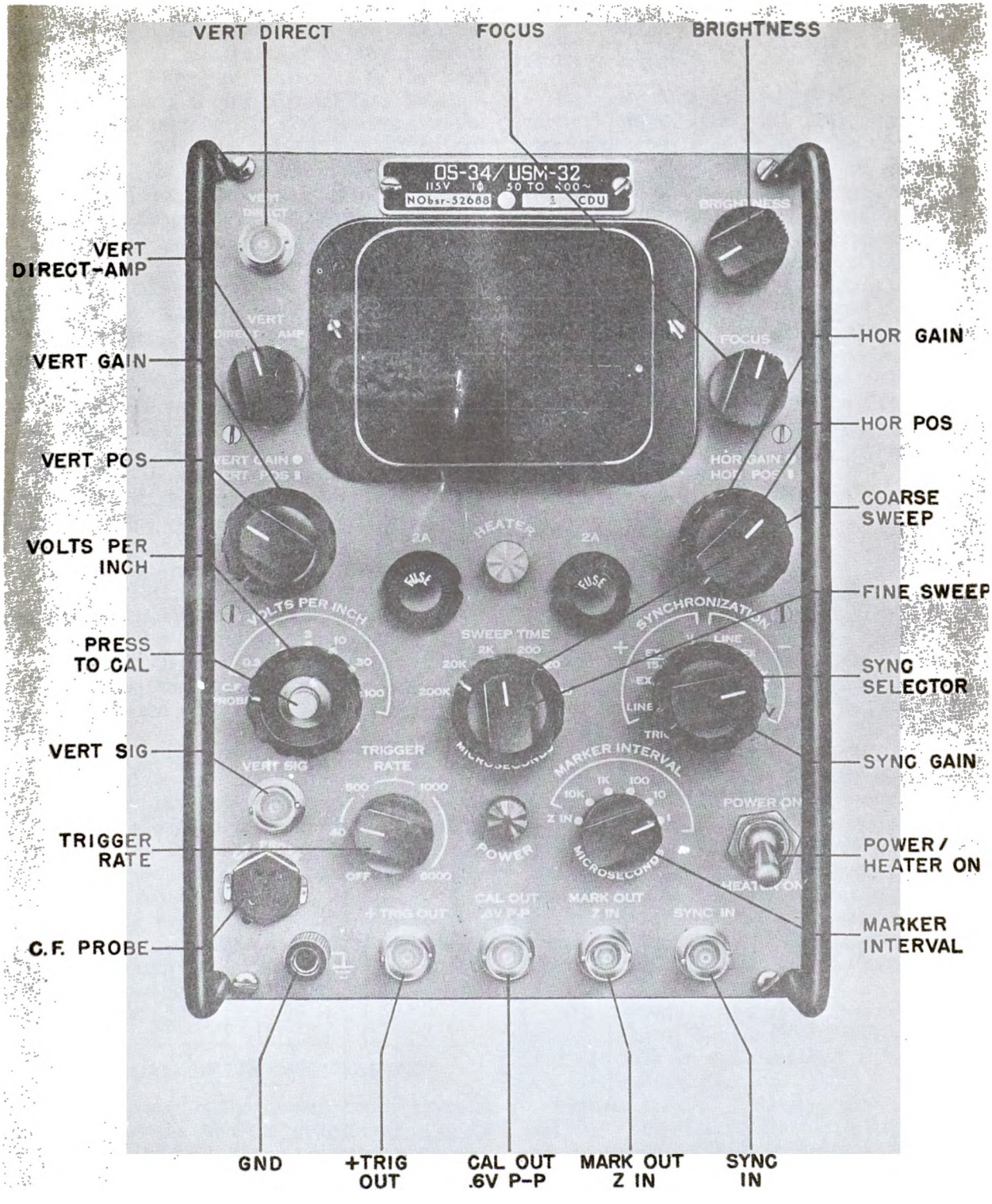

Figure 17-15.—Front panel of the oscilloscope AN/USM-32.

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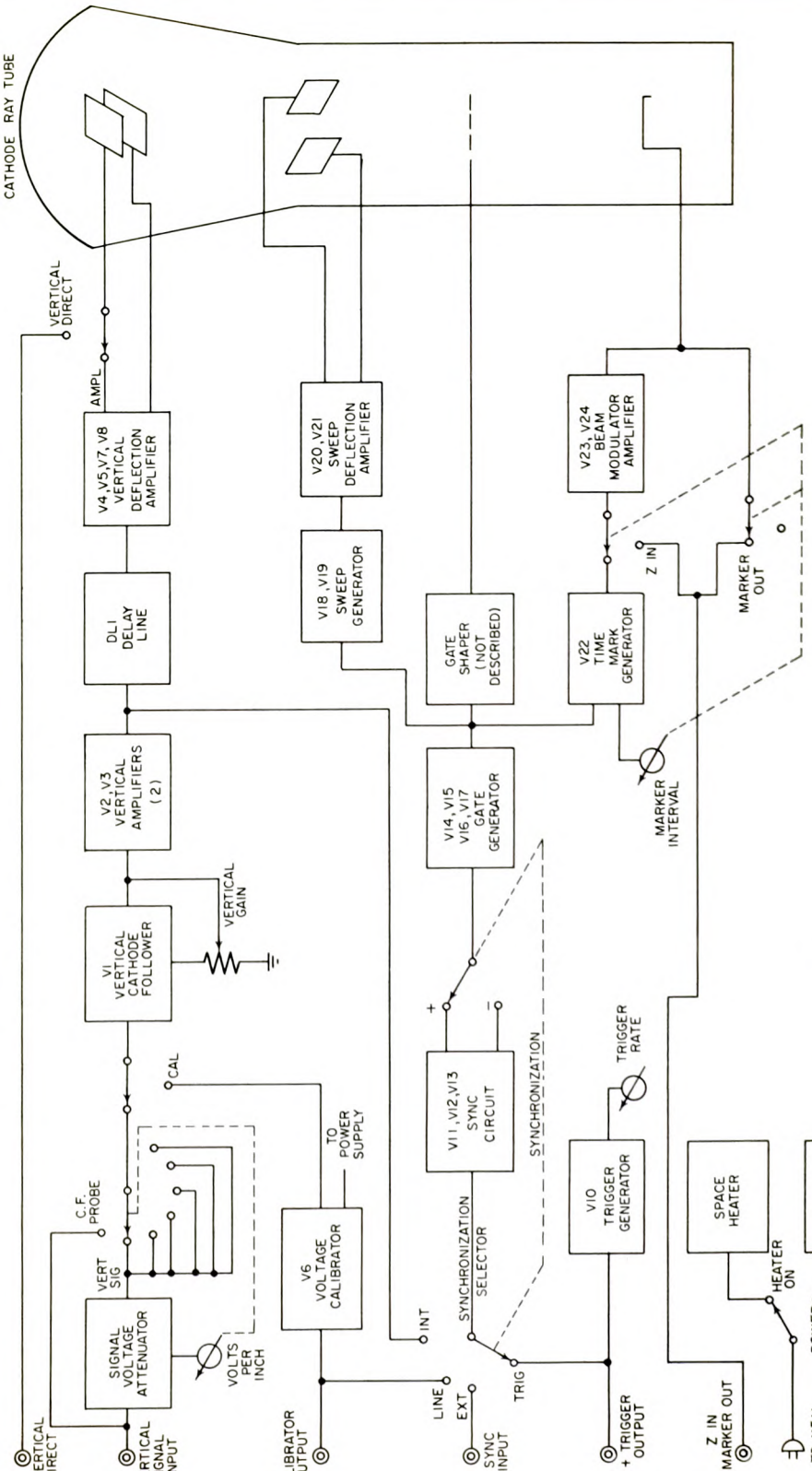

Figure 17-16.—Synchroscope block diagram.

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front panel. The markers are available at the cathode-ray tube cathode, and at the MARK-OUT/Z-in terminal for external use when the marker interval control is in any position except Z-in. When the control is in the Z-in position, external timing signals may be connected to this terminal to give markers on the sweep trace. Positive pulses applied to the MARK-OUT/Z-in terminal will produce blanking markers on the oscilloscope screen, and negative pulses will produce brightening markers.

The trigger generator supplies the internal trigger pulse for the sweep circuit when the synchronization control is in the TRIGGER position (as shown). The signal is also available at the +trigger-out terminal for external test purposes. The duration of the output trigger pulse is approximately two microseconds, and has a repetition rate that ranges from 45 cycles to 5.5 kilocycles, as determined by the setting of the trigger rate control.

The built-in voltage calibrator provides a method of making quantitative amplitude measurements of the vertical input signal. The circuit produces a square-wave calibrator signal, which is available internally, and may be substituted for the vertical input signal by depressing the press-to-cal pushbutton on the front panel. In this manner, the unknown signal can be compared with the known signal (calibrator) voltage to determine its amplitude. The signal is available at the calibrator output terminal at all times.

Regulated potentials of +110 v and -87 v, as well as +310 v and -1700 v unregulated, are obtained from the power supply for use in the circuits of the oscilloscope. A 6.3-v, a-c source is contained in the power supply for servicing the various tube heaters and the power indicator lamps on the front panel.

The power supply circuit is a conventional electronically controlled source. The operation of these circuits is considered in Chapter 3 of *Basic Electronics*, NavPers 10087, and is not repeated here.

VERTICAL DEFLECTION CIRCUIT

The discussion of the circuitry of the vertical deflection system, and the synchroscope in its entirety, is limited to those circuits associated with the operating controls. A simplified schematic of the vertical deflection system is shown in figure 17-17. The primary function of the vertical deflection system is the amplification of the vertical signal input before it is applied to the vertical plates of the cathode-ray tube.

Vertical Controls

Input signals are normally applied to the vertical signal jack, J1. The signal voltage attenuator (operated by the volts-per-inch control) is R-C compensated to maintain frequency and phase response over the range of the vertical amplifier. The attenuator, regardless of the switch position, presents an impedance of one megohm in shunt with approximately 28 $\mu\mu$f to the circuit under test. The stray capacitance from the input terminal to ground is considered in this total. If the vertical amplifier sensitivity is set for 0.3 v peak-to-peak per inch by means of the vertical gain control and built-in voltage calibrator, the switch settings (as labeled on the panel) determine the voltage that must be applied to the vertical signal terminal to give one inch deflection.

Cathode Follower Probe

The volts-per-inch control (fig. 17-17) is shown in the C. F. PROBE position. A schematic of the cathode follower probe circuit is shown in figure 17-18.

The cathode follower probe is employed where a very high impedance must be presented to the circuit under test with negligible loss in the oscilloscope sensitivity. With switch SID (fig. 17-17) in the C. F. PROBE position (as shown), power is supplied to the cathode follower probe connector from the 310-v supply. R13, in series with the 310-v source, permits the application of 110 volts to pin E of the connector. This potential is applied through pin E of P3 (fig. 17-18) to V9, which is connected as a triode and functions as a cathode follower. R66 is a parasitic suppressor. C31 isolates the cathode follower from d-c potentials that may be present at the input tip.

The output of the cathode follower is developed across R68 and R69 and coupled through peaking coil L8 and terminal A to the volts-per-inch switch, S1A (fig. 17-17).

Signals from J1 (except in the C. F. PROBE and 0.3 v position of S1) are developed across one of the attenuator voltage dividers as determined by the setting of S1A. Output signals from the attenuator are taken from a tap on the divider and coupled through S1C and the press-to-cal button to the input circuit of V1. In the

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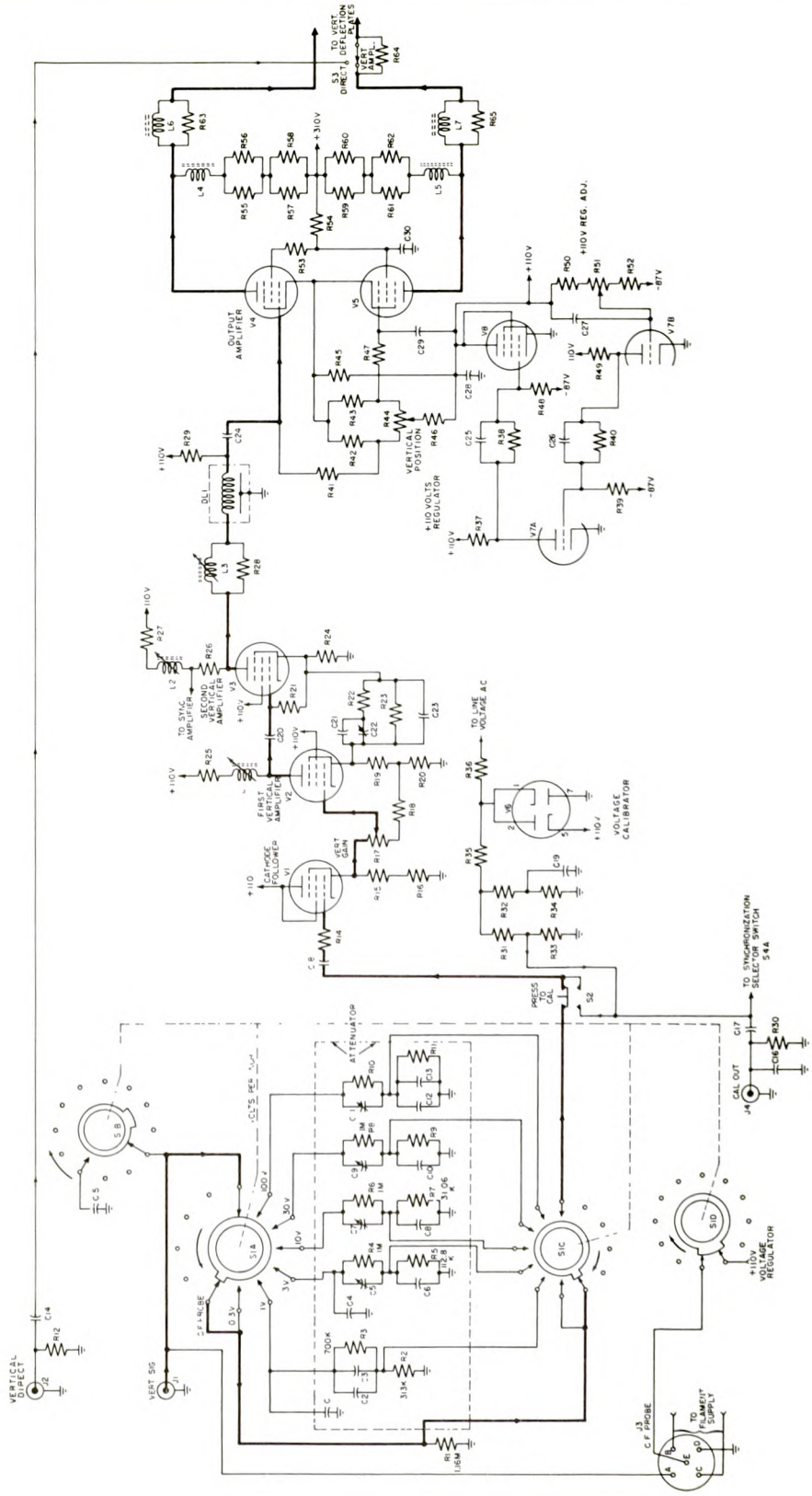

Figure 17-17.—Vertical deflection circuit.

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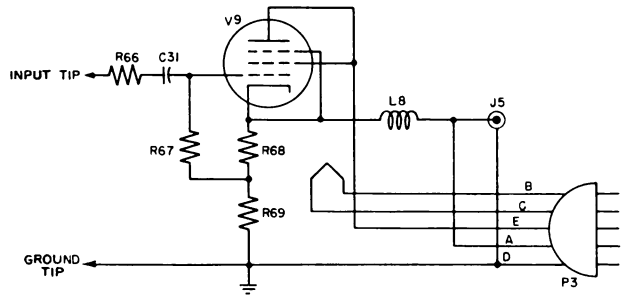

Figure 17-18.—Cathode follower probe.

C. F. PROBE and 0.3 v position of S1, the attenuator is bypassed, and R1 develops the signal input.

Input Cathode Follower

Cathode follower, V1 (fig. 17-17) provides a high impedance to the circuit under test in all positions of the volts-per-inch switch, S1. C18 is a coupling capacitor to the V1 grid. R14 acts as a parasitic suppressor.

The vertical gain control is connected between the cathodes of V1 and V2 in such a manner that both ends of the gain control are near the same d-c potential. This permits varying the amplitude of the signal without changing its relative position on the screen. The gain control, when used in conjunction with the input attenuator, provides continuous variation in the amplitude of the input so that signals from 0.1 to 300 v will produce one-inch deflection on the screen of the cathode-ray tube.

Delay Line

An artificial delay line, DL1, in the plate circuit of V3 delays the signal 0.35μ with respect to the signal coupled to the sync amplifier (V11 of figure 17-19). L3 and R28 form a damped series peaking network at the delay line input. These components, in conjunction with R29, make up a part of the delay line termination impedance.

Output Amplifier

The cathode-coupled, vertical-deflection amplifier (V4 and V5) converts the unbalanced vertical input to a balanced push-pull output.

The d-c potential at the grids and cathodes of V4 and V5 is taken from the junction of R46 and R50. The signal voltage developed across the common cathode resistor, R45, results in an effective signal at the control grid of V5. The signal thus obtained at the V5 grid is approximately equal in amplitude and opposite in phase to that at the V4 grid. The resulting signal at the plate of V5 is approximately equal in amplitude and opposite in phase to that of V4, and is coupled to the other vertical deflection plate.

The d-c potential at the grids of V4 and V5 is controlled by the position of the arm of vertical position control, R44. Moving this control changes the grid bias potentials of the tubes. The resulting change in the plate potential of V4 and V5 re-positions the pattern on the screen.

When the vertical-direct amplifier switch, S3 (in the plate output circuit) is set to the VERT-AMP position, it connects the amplifier output to the vertical deflection plates of the cathode-ray tube. When S3 is set to the Direct position, the input to the vertical amplifiers is disconnected, and the signal input connections must be made at the vertical direct terminal. This permits connection of the signal directly to the vertical deflection plates. Vertical positioning is still available because the vertical amplifiers are connected to the vertical deflection plates through R64. R12 is the terminating resistor for the vertical-direct input, which is shunted by the stray capacitance.

Voltage Calibrator

Accurate amplitude measurements of the vertical signal input at J1 may be made through the use of the built-in voltage calibrator, V6. The calibrator signal may be substituted for the vertical signal by depressing the press-to-cal pushbutton.

The power line voltage is applied to V6 through limiting resistor R36. On the positive half cycle, the voltage is clipped to 110-v peak, due to the conduction through V6 from cathode (pin 5) to plate (pin 2) when the plate voltage exceeds the +110 v applied to its cathode as bias voltage. Since the pin 7 plate is tied to ground, any negative voltage impressed on this section of the tube will cause it to conduct, thus clipping the negative half cycle at approximately ground potential. The square-wave voltage at the pin 2 plate of 110-v peak is developed across R35 and two voltage dividers. The voltage dividers comprise R31 and R33, and R32 and R34 respectively. The voltage division is such

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that 1.2 v-peak exists across R33. The 1.2-v peak signal is also developed across C17.

On the positive half cycle of applied voltage, C17 charges to a 1.2-v peak. On the negative half cycle the voltage across C17 is divided equally between R30 and R33. Thus, the voltage across R33 varies between a +1.2-v peak and a +0.6-v peak with respect to ground for each cycle of applied voltage. In this manner, when the press-to-cal button is depressed, 0.6 v peak-to-peak is applied to the vertical cathode follower, V1. The calibrator output voltage is also available at the CAL-out jack, J4.

HORIZONTAL DEFLECTION CIRCUIT

The horizontal deflection (fig. 17-19) provides the necessary signals to the horizontal deflection plates to produce the required sweep trace and blanking. The synchronization-selector, sweep-time, horizontal-gain, and horizontal-positioning controls permit the proper regulation of the sweep signals.

Synchronization Selector

The synchronization selector switch, S4, permits the operator to choose the method and polarity of sweep initiation. In the ± line position of the switch, the trigger signal is derived from the output of the voltage calibrator, which, in turn, was derived from the line voltage. The ± external position of S4 permits the sweep triggering to be applied from an external source connected to the sync in terminal, J7. The external signal is attenuated 15:1 by the voltage divider comprising R77 and R78 when the synchronization selector is placed in the ± EXT/15 position. The ±V positions of S4B couple a portion of the undelayed signal from the vertical amplifier to the sync amplifier to activate the sweep circuit. Since this input is taken from the vertical amplifier (V3 in figure 17-17) ahead of the delay line, the sweep trace will appear 0.35 μs before the vertical signal that initiated the sweep. Likewise, the delay is only available when the synchronization selector (fig. 17-15) is in the ±V positions. There is also a trigger position available on S4, which allows a portion of the positive internal trigger generator signal to trigger the sync circuit.

The selected sync signal is coupled through C34 to the control grid of the first sync amplifier, V11. The sync gain control, R81, in the cathode of V11 allows for variation in the amplitude of the V11 plate output.

Sweep circuits of oscilloscopes may be driven, free-running, or both. Free-running oscilloscopes usually make use of a portion of the vertical input signal to synchronize a free-running sawtooth generator. Driven oscilloscopes (synchroscopes) require a triggering signal from some source (either internal or external) to initiate the cathode-ray sweep.

The sweep circuit in figure 17-19 employs a driven-type sweep circuit only, and the cathode-ray beam is blanked, except during the sweep period. No trace will appear until the synchronizing signal is sufficient in amplitude to trigger the sweep circuit.

Because the gate generator multivibrator, V14 and V15, requires a negative pulse input, and a phase reversal takes place in the second sync amplifier, V13, the input to V13 must be a positive signal. The circuit of V12 is connected as a phase splitter. In the ± position (as shown) V13 receives the output from the plate of V12. If the sync input to V11 is a positive signal, the input to V13 will also be positive. If the input to V11 is a negative signal, S4B must be in the - position, and V13 receives the output from the cathode of V12, which will be a positive signal.

Sweep Controls

The chosen coarse sweep capacitor, C47 through C50, which is selected by S6 and connected from the plate to the cathode of V18, is effectively discharged as a result of the V18 conduction. A negative gate from V14 cuts off V18, which permits the charging of the selected coarse sweep capacitor through R119, and R120 from the +310-v supply. R115 is the sweep time (fine) control, which provides continuous adjustment of the sweep duration.

When the highest sweep range is chosen, Cd, which represents the stray capacitance of the circuit, is selected by S6. In addition, R115 is shunted by R106, which lowers the total resistance of the circuit and lowers the charge time.

In operation, first set the sweep time (coarse) control to the desired position. The sweep time (fine) control should then be advanced until a satisfactory pattern is displayed on the screen.

An approximation of the sweep period may be obtained by reading the indication on the sweep controls. However, a more accurate

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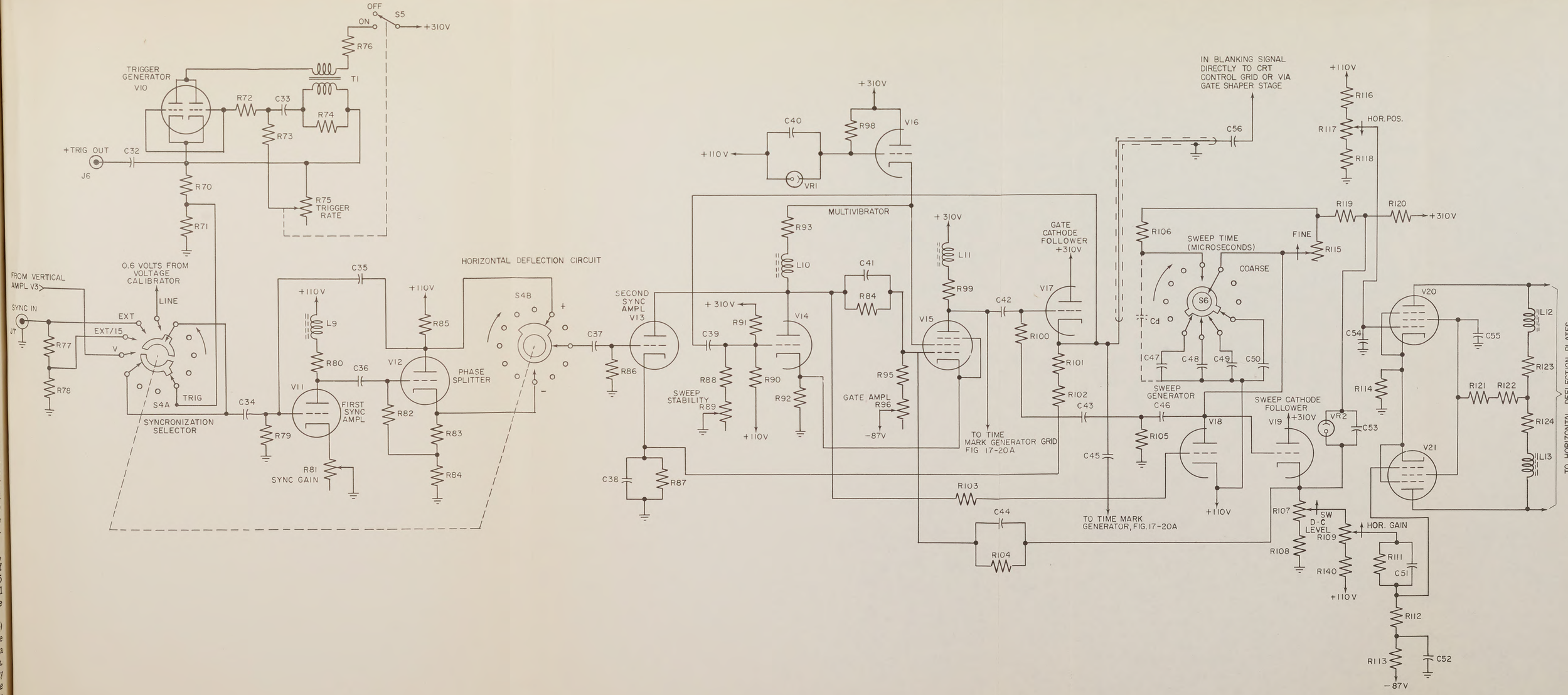

Figure 17-19.—Horizontal deflection circuit.

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method of measuring sweep duration (period) can be obtained through the use of the built-in time calibrator discussed later.

A portion of the sawtooth voltage appearing across the sweep-output cathode-follower (V19) load resistors is applied to the horizontal gain control, R109. One end of R109 is connected to the adjustable arm of the sweep d-c level control, R107. This control sets the d-c level at the control grid of the output tube, V21, and when properly adjusted, it provides an equal sweep trace expansion from both sides of the center of the cathode-ray screen when the horizontal gain is advanced.

The trigger generator stage comprising tube V10 provides a source of positive triggers for initiating the action of the sweep circuits at variable repetition rates. The stage functions as a standard blocking oscillator with positive feedback supplied by T1. Trigger rate control R75 varies the grid time constant and likewise the rate of the pulses from the trigger generator. S5 completes the path from the B supply to V10 when in the ON position.

The trigger output is taken from the cathode of V10. The positive pulses have an amplitude of 100-v peak from cathode to ground, and are fed to the +trig-out jack, J1. The portion of the pulse developed across R71 is applied to the synchronization selector switch for sweep initiation.

The trigger repetition rate is continuously variable from 45 to 5500 PPS. The trigger generator can be disabled by rotating the trigger rate control to the OFF position, which opens S5.

TIME CALIBRATION CIRCUIT

Measurement of pulse width, or determination of frequency (sweep writing rates) is accomplished by superimposing blanking markers on the desired signal. The marker interval selector switch, S7 (fig. 17-20) determines the time interval between successive blanking or timing markers, and is calibrated in microseconds. Intervals of 1, 10, 100, 1,000, 10,000 μs may be selected. The switch is shown in the 100-μs position.

To use the time calibration provided, rotate the marker interval switch, S7, to the appropriate position, whereupon the timing markers will be presented as blanking markers or dark spots (positive input) along the length of the trace. The interval between them will be that indicated by the switch setting with a time accuracy of ± 5%. Determine the period of one cycle of the input signal and compute the frequency by using the formula,

$$F(\text{frequency}) = \frac{1}{P(\text{period})}.$$

The output from the time mark generator, V22, is taken from R135 to ground and applied through S7D and C70 to the grid of the first beam modulator, V23. The internal modulating signals from the second amplifier, V24, are applied to the cathode-ray tube.

When S7 is placed in any position other than Z in, the internally generated timing markers are coupled to the front panel jack, J3, labeled, mark out-Z in. The jack is shown as J3 in figure 17-20, A. S7 is shown in the Z in position, in figure 17-20, B. With the switch in this position, external modulation signals are coupled through S7C, S7D, and the beam modulator amplifiers to the cathode-ray tube. The internally generated marker output from R135 in figure 17-20, A, (time mark generator) to the beam modulator amplifiers is disconnected by S7D.

Positive peaks arriving at the cathode of the cathode-ray tube with a magnitude of 10-v peak will cause blanking markers on the cathode-ray screen. Negative signals of sufficient magnitude cause brightened markers to appear.

CATHODE-RAY TUBE CIRCUIT

Negative, high voltage for the cathode-ray tube is obtained from a -HV supply and applied across a voltage divider (fig. 17-21), which develops the proper operating potentials for the tube. A voltage regulator tube, VR3, maintains a part of this voltage at -87 v with respect to ground, which is used as a fixed bias potential in the circuits of the oscilloscope. The potential for the focus and brightness controls is also obtained from this divider. The astigmatism control receives its potential from a voltage divider connected from the +310-v supply to ground.

A sharply focused line or spot of high intensity, having short length or small area, should not be permitted to remain stationary on the screen for any considerable length of time. Under such conditions, the entire energy of the beam is concentrated over a small area, thus subjecting the screen material to burning and discoloration.

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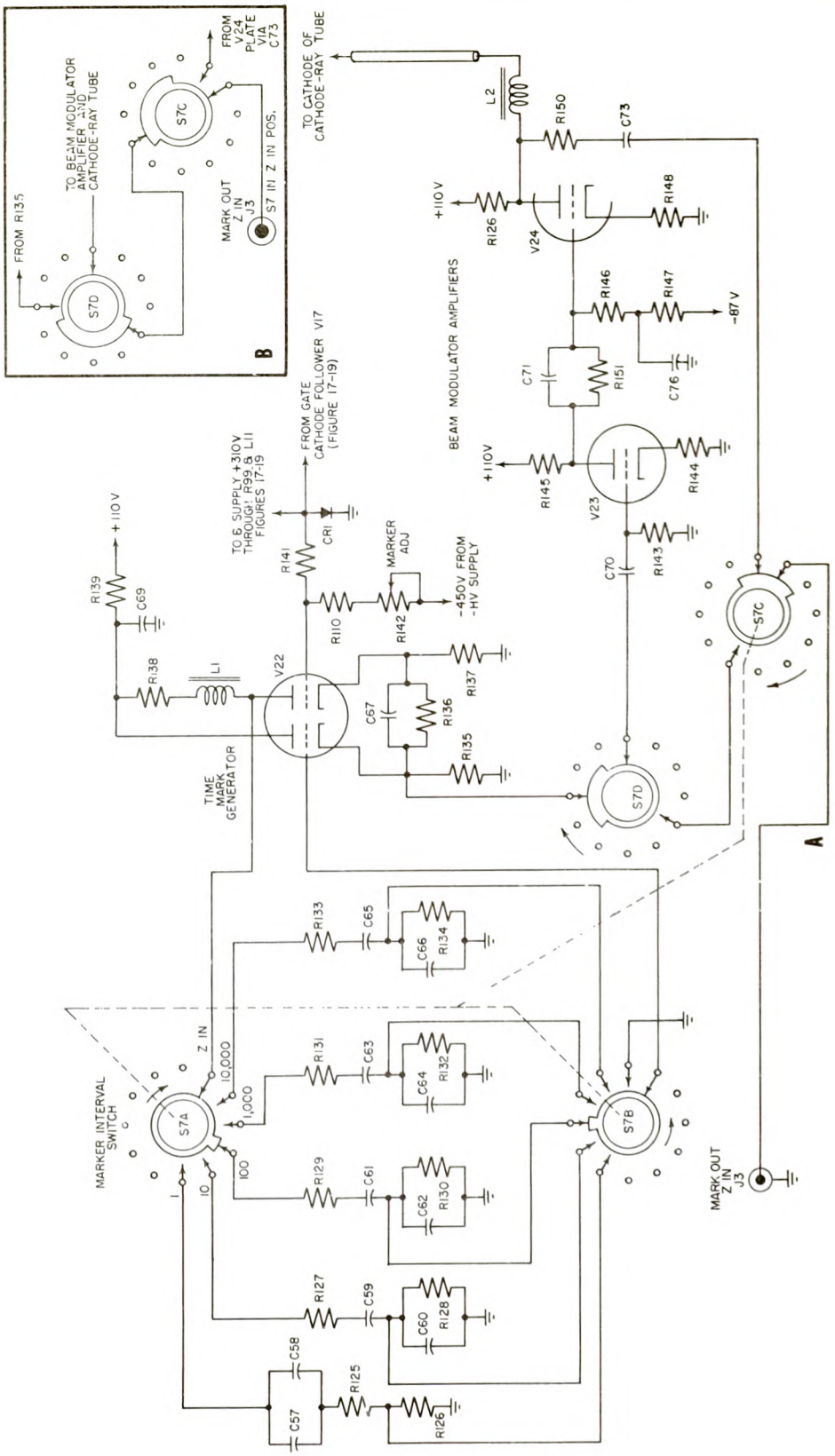

Figure 17-20.—Time calibration circuit.

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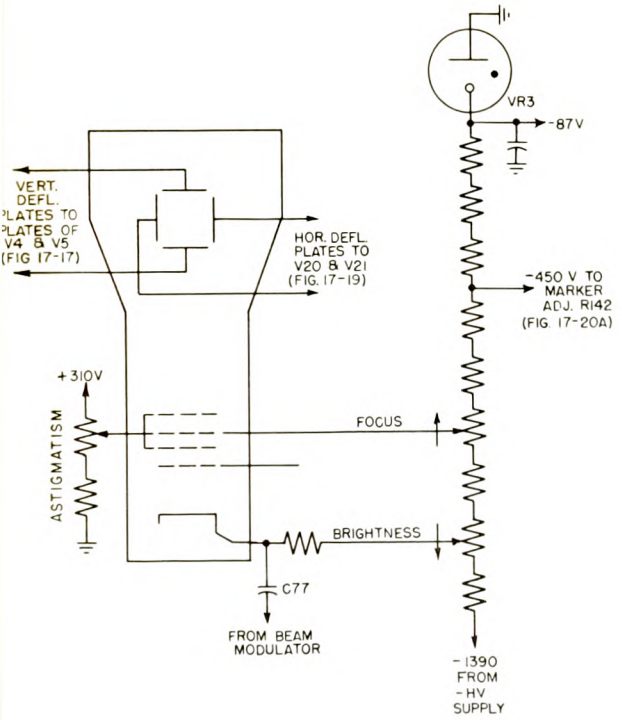

Figure 17-21.—Cathode-ray tube circuit.

SIGNAL MEASUREMENTS

Amplitude Calibration

Amplitude measurements of the vertical signal may be made through the use of the built-in voltage calibrator, as previously discussed. The signal from the voltage calibrator is substituted for the input signal by depressing the press-to-cal pushbutton.

When calibrating the oscilloscope, the vertical gain control must be set to its fully clockwise position. Thus, the position of the volts-per-inch control alone will affect the magnitude of the signal fed to the vertical amplifier stages.

Since the frequency of the calibrating voltage is the same as the line frequency (as discussed) the synchronization selector switch should be placed to either its + or - LINE positions. The 20K-2K position of the sweep time (coarse) control may be used. The position of this control is arbitrary since it only determines the number of cycles appearing on the cathode-ray screen. The synchronization (gain) control determines the magnitude of the synchronizing voltage, and must be advanced sufficiently clockwise to trigger the sweep circuits.

The press-to-cal pushbutton should now be depressed and the sweep time (fine) control adjusted for several cycles of the calibration pattern. Set the vertical gain control to give two inches of deflection on the cathode-ray screen. Do not change the position of the vertical gain control after it has been set for signal amplitude measurements. If it is accidentally moved, it must be reset through the same procedure.

The peak-to-peak voltage connected to the vertical signal terminal may now be determined. First, note the signal deflection in terms of inches on the cathode-ray screen. Secondly, multiply this reading by the volts-per-inch selector switch setting. For example, if the volts-per-inch switch is set to the 3 position, a deflection of 1.5 in. will represent 4.5 v.

The screen may be calibrated for other values of volts-per-inch than those indicated on the volts-per-inch selector switch. If the vertical gain control is adjusted to give one inch of calibrator square-wave deflection, the scale will be calibrated for 0.6, 2, 6, 20, 60, or 200 volts-per-inch at the respective volts-per-inch selector settings.

Use of Test Probes

The cathode follower probe (fig. 17-22) is employed where an especially high impedance and low shunting capacity must be presented to critical circuits being tested. The probe attenuates the input signal approximately 30%.

To make quantitative measurements when the cathode follower probe is being used, use special calibration procedures as follows: touch the probe tip to the CAL-out terminal (fig. 17-15), which provides a 0.6-v peak-to-peak signal. Adjust the vertical gain control for 0.6-v, peak-to-peak, full-scale (2 in.) deflection.

A maximum signal of 2 volts rms may be investigated by the cathode follower probe without distortion. The 2-volt signal may be superimposed on a d-c level as high as 600 volts.

The attenuator probe (fig. 17-23, A) is employed where negligible loading of a high impedance circuit is important, and where sufficient signal amplitude is present to allow an attenuation of the impressed signal by 20:1. With the same impedance requirements, the attenuator probe should be employed in those applications where the cathode follower probe would not be suitable.

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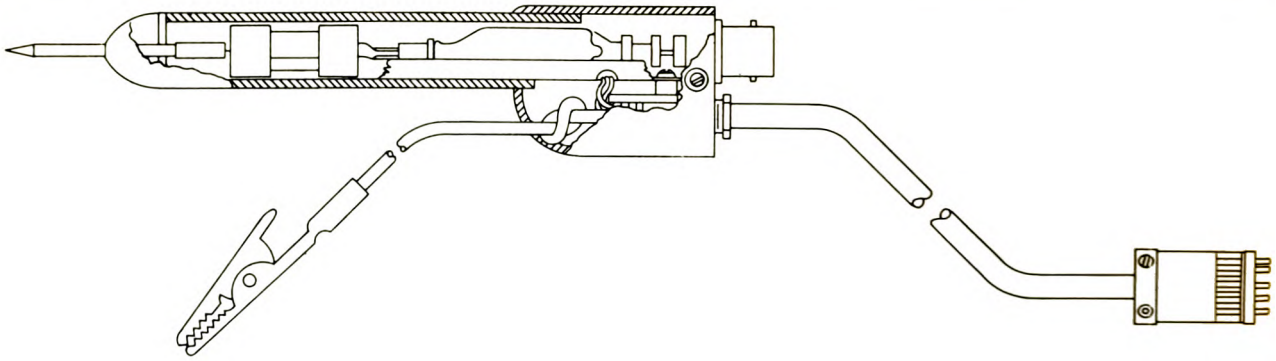

Figure 17-22.—Cathode follower probe, pictorial diagram.

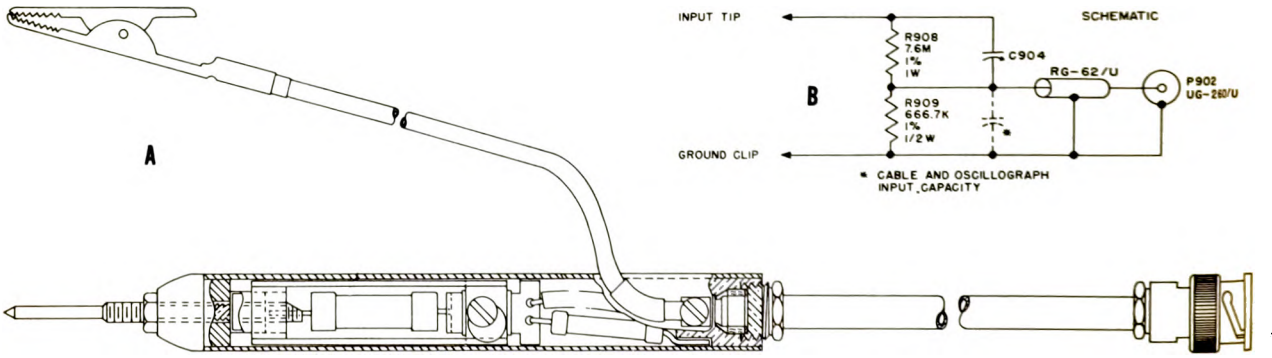

Figure 17-23.—Attenuator test probe.

The maximum permissible a-c signal input for the attenuator probe is 600 v peak-to-peak. Any signal coupled to the probe is attenuated 20:1, and this ratio should be kept in mind when the built-in voltage calibrator is employed for amplitude measurements.

The signal cable for the attenuator probe is permanently attached to the probe since the cable capacity (fig. 17-23, B) forms part of the frequency compensation circuit of the probe. For normal signal input applications, the probe is connected to the vertical signal terminal. However, it may also be connected to the vertical direct, the mark out Z in, or the sync-in terminals.

The detector probe (fig. 17-24, A) will demodulate a-m signals in the range from 10 cycles to 15 kc over a carrier frequency range of 0.5 to 400 mc. The audio frequency response is within 6 db at any carrier frequency in the operating range.

The peak inverse voltage of the crystal (fig. 17-24, B) is 125 v. Signal voltages to be demodulated by the detector probe should not exceed this value.

To use the detector probe, connect one end of the r-f cable (provided with the equipment) to the vertical signal terminal and the other to the probe. Apply the probe tip to the circuit under test.

The resistor assembly (fig. 17-25, A) is a plug-in adapter that connects to the cathode follower probe socket. When plugged in, it provides 75-ohm termination (fig. 17-25, B) for the vertical signal terminal in all positions of the volts-per-inch switch.

Adapter Connector

The adapter connector (not shown) enables the operator to connect wire-type test leads to the coaxial terminals on the front panel. Set knobs are provided on the adapter for tightening the wire leads.

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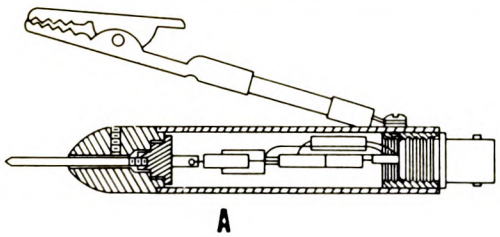


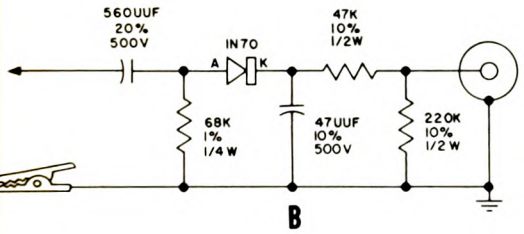
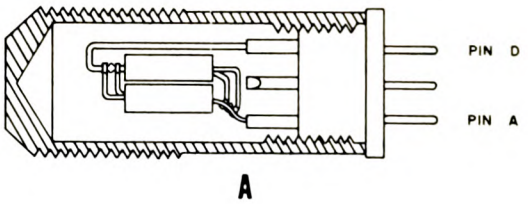


Figure 17-24.—Detector test probe.



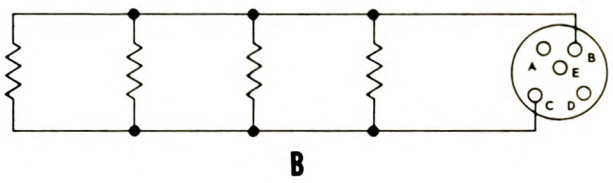


Figure 17-25.—Resistor assembly.

QUIZ

1. The radio interference measuring set AN/URM-47 is a highly sensitive h-f, v-h-f, and u-h-f superheterodyne receiver covering what frequency range?
2. When the calibrator is on, what major adjustment can be made?
3. The demodulated signal from the detector (fig. 17-2) is acted upon by the meter detector weighting (measuring) circuits and applied to what stage?
4. The visual null indicator (fig. 17-2) is provided primarily for use in what type calibration?
5. F-m signals may be received with reasonable audio quality through the use of what type detection?
6. Because of the wide frequency range covered by continuous and discontinuous interference, the magnitude of the interference will depend upon what feature of the measuring equipment?
7. Field intensity is the value of the electric field at a given point, and is measured in what unit?
8. In the CAL position of the attenuator (fig. 17-11) the input signal is terminated in what type load?
9. The input filter Z4 (fig. 17-11) is a transmission line, low-pass filter for attenuating frequencies in what range?
10. In the X1 position of the r-f attenuator (fig. 17-12) the output from the secondary of T6 is delivered to the control grid of the first i-f amplifier through what contacts of relay K1?
11. A calibrate control is connected in the cathode circuit of V8 (fig. 17-12) for what purpose?
12. The fourth i-f stage, V11 (fig. 17-12) is an equalizing or gain compensating stage having its gain controlled in what manner?
13. Function switch section S1B (fig. 17-12) provides for application of +225 v to the plate of the pulse generator (V7 of figure 17-11) so that the tube operates and actuates what circuit?
14. Both sections of the detector diode (fig. 17-12) operate when the function switch, S1, is placed in what position?
15. The efficiency of the detector diode (fig. 17-12) in quasi peak operation differs from that in field intensity for what reason?
16. Tube V17 (fig. 17-13) is connected in what type circuit?
17. The vacuum tube voltmeter stage (fig. 17-13) comprises what type stage?
18. Output signals from the calibrating circuit (fig. 17-11) are coupled to the RI-FI meter input when the function switch is set to what position?
19. Input signals from the vertical signal terminal (fig. 17-15) are first applied to what circuit?
20. Timing marker intervals of 1, 10, 100, 1000, and 10,000 μs may be selected by what control on the oscilloscope front panel (fig. 17-15)?

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21. What circuit (fig. 17-16) supplies the internal trigger pulse for the sweep circuit when the synchronization switch is in the TRIGGER position?
22. The cathode follower probe (fig. 17-18) is employed for what purpose?
23. An artificial delay line DL1 (fig. 17-17) in the plate circuit of V3, provides what amount of vertical signal delay with respect to the signal coupled to the sync amplifier?
24. In the LINE position of the synchronization switch (fig. 17-19) the trigger signal is derived from the output of what circuit?
25. When the highest sweep range is chose (fig. 17-19), what capacitor is selected b S6?
26. The trigger repetition rate of the trigge generator (fig. 17-19) is continuously vari able between what limits?
27. Positive peaks arriving at the cathode o the cathode-ray tube with a magnitude o 10 v peak will have what effect on th cathode-ray screen?

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CHAPTER 18

THE USE OF SPECIAL TEST EQUIPMENT, PART II

SPECTRUM ANALYZER

A spectrum analyzer is a device that sweeps ver a band of frequencies to determine (1) what requencies are being produced by a specific ircuit under test and (2) the amplitude of each requency component. To accomplish this, the pectrum analyzer presents, on an oscilloscope isplay, a pattern in which the relative ampli- udes of the various frequencies of the spectrum re plotted on the vertical, or Y axis, while the requencies themselves are plotted on the hori- ontal, or X axis, of the cathode-ray tube. The ver-all pattern of this display indicates the roportion of power present at the various fre- uencies within the spectrum. A representative pectrum analyzer equipment is the AN/UPM-3, which consists of the TS-148/UP Spectrum nalyzer with carrying case and accessories.

The control panel of the TS-148/UP spec- rum analyzer is shown in figure 18-1. The cir- uits of this unit will be discussed in this section ince they represent to a great extent spectrum nalyzer circuits in general.

The spectrum analyzer can be used to ex- mine the spectra of magnetrons, local oscilla- rs, test sets, and other equipment operating ithin its frequency range. A frequency meter s used to measure accurately the frequencies any of the above equipments, or to set the equency of radar and beacon local oscillators radar sets. The spectrum analyzer can also e used as a f-m oscillator to tune T/R boxes nd ATR boxes in transmitters.

COMPONENTS OF A SPECTRUM

A common conception of the output of a pulsed scillator is a single frequency, which is turned and off for periods of standard duration. This tion is similar to the output of a conventional w telegraph transmitter. The output of the ulsed radar oscillator does not consist of a ndamental frequency that is turned on and off, but must be considered as a fundamental frequency that is pulse modulated by the waveform of the trigger pulse.

Effects of Modulating Pulses

When modulated, any fundamental frequency will produce a fundamental frequency with side-band frequencies, which collectively is called a spectrum. The distribution of the power on these frequencies is a function of the modulation. Normally, modulation is plotted on an amplitude and time basis as shown in figure 18-2, B, C, and D.

Assume F to be the fundamental frequency of the oscillator. Waveform A shows the fundamental frequency plotted as amplitude against time. The number of periods occurring within one second determines the frequency of the oscillation. The amplitude is represented as proportional to the distance between the negative and positive peak of one cycle.

In a spectrum pattern, this same frequency and amplitude would be represented as F in figure 18-2, E. Points along the horizontal axis represent frequency, which increases from left to right, while distance along the vertical axis above the baseline represents amplitude. Thus, two methods of diagramming the results of amplitude-modulating a carrier frequency are shown.

Assume that F_p1 is a modulation frequency applied to the fundamental frequency, F. This is normally represented on an amplitude-versus-time basis (fig. 18-2, B). This same type of modulation is shown by two lines on the spectrum pattern. These lines are marked F+Fp1 and F-Fp1 in figure 18-2, E. The reason for this is that the modulated wave actually represents the results of heterodyning two different frequencies. These frequencies are effectively present in the modulated output, and can be detected by suitable receivers. One of

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Figure 18-1.—Spectrum analyzer control panel.

the frequencies is the sum of the two, and the other is the difference. The amplitudes of the new frequencies are each half the amplitude of the modulating frequency.

Assume also that a second harmonic of the modulating frequency exists. This is usually of a smaller amplitude than the fundamental of the modulating frequency. Another set of waves will be developed as at C for amplitude-versus-time, and at F + Fp2 and F - Fp2 on the spectrum pattern.

Additional modulating frequencies will produce additional sideband frequencies. Since these frequencies are normally present in a harmonic relationship, the net result is a number of different frequencies above and below the carrier. The difference between any two adjacent frequencies is equal to the fundamental modulating frequency.

Spectrum of a Pulsed Oscillator

Usually a pulse modulated oscillator is pulsed by the application of a rectangular wave of voltage to an oscillator circuit. A narrow rectangular wave contains an exceedingly wide range of harmonics, including harmonics of a very high order. Consequently, a pulsed oscillator may be assumed to be an oscillator modulated by a modulation frequency that is exceedingly rich in harmonics.

The pulse frequency and the basic frequency of the pulsed oscillator are shown on an amplitude-versus-time graph (fig. 18-3). From this figure it might appear that the output is simply a pure c-w wave of constant amplitude and frequency, turned on for brief intervals of time. This concept cannot be true because of the presence of the modulation frequency Fp1, and the very large number of harmonics (of the modulating frequency) Fp2, Fp3, etc., as discussed. Therefore, the fundamental modulation frequency and its many harmonics may be considered to modulate the oscillator to produce a fundamental frequency with many sidebands. The net result is a spectrum such as was developed in figure 18-2, E, except that it is extended to a very large number of frequencies above and below the fundamental

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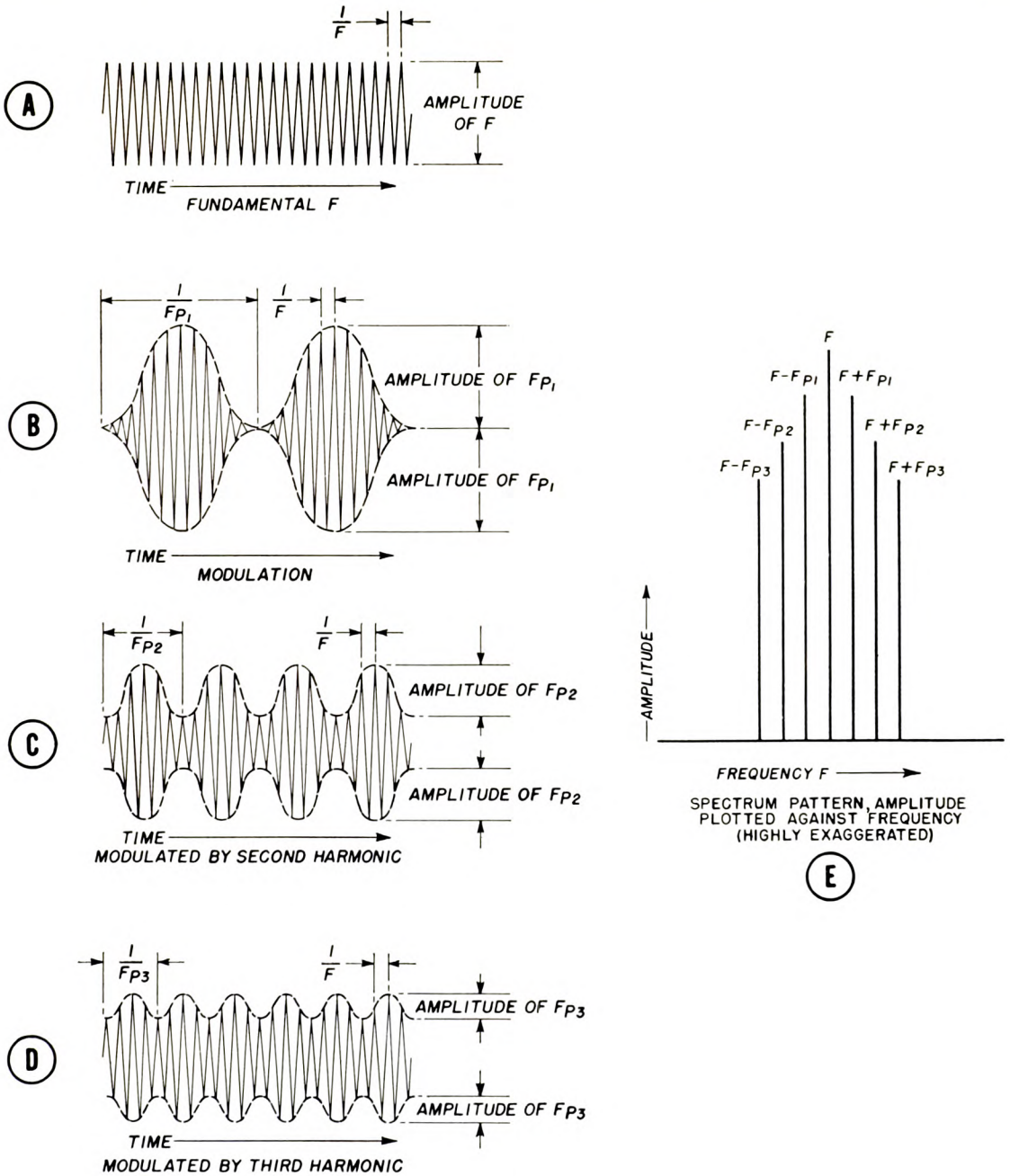

Figure 18-2.—Comparison of amplitude vs time to amplitude vs frequency.

frequency. Such a spectrum is shown in figure 18-4, A, along with the pulse that produced the spectrum in figure 18-4, B. It is shown that the output of the oscillator consists of an infinite number of lines representing different frequencies. However, because of the harmonic relationship between the modulation frequency and its harmonics, these lines will always be separated by a distance on the baseline that is equal to the fundamental modulation frequency, Fp1.

The amplitude-versus-frequency plot provides an envelope, which is of value in estimating the power distribution in the output of

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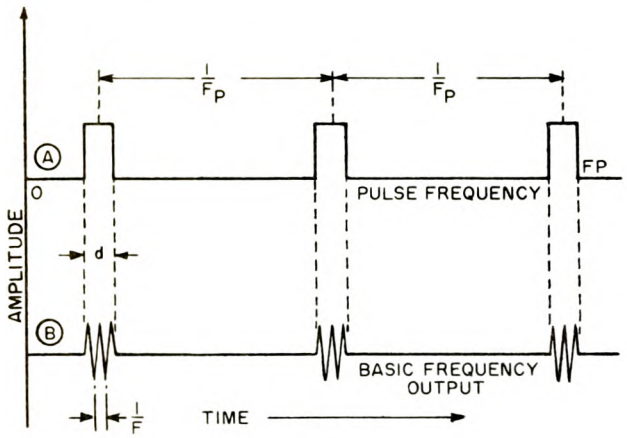

Figure 18-3.—Output of a pulsed oscillator.

a pulsed oscillator. This spectrum, as plotted on the spectrum analyzer, is a power spectrum produced by the square law characteristics of the detector discussed later. The true power spectrum, which represents the voltage (or amplitude) squared (E^2), is shown in figure 18-4, D.

The power spectrum emphasizes the importance of confining the majority of the power at the fundamental frequency of the pulsed oscillator. The spectrum analyzer can be used in tuning up a pulsed oscillator so as to provide the greatest range of power output in the bandpass circuits of a receiver.

Figure 18-4, C, represents a sample of the spectrum (fig. 18-4, A). The figure shows the effective voltage derived from the oscillator frequency spectrum. We can conclude that the pulsed oscillator output contains both frequency and amplitude components. The effect of amplitude modulation is to increase the number of sidebands in the spectrum. The effect of frequency modulation is to increase the amplitude of the side lobes, since the frequency of the oscillation is effectively shifted back and forth between points on each side of its true fundamental frequency.

BLOCK DIAGRAM

Basically, the spectrum analyzer consists of a superheterodyne receiver with a frequency-modulated, r-f oscillator (fig. 18-5). The analyzer r-f oscillator is modulated with a sawtooth voltage, which causes the oscillator to operate over a range of frequencies. As the frequency of the analyzer r-f oscillator increases, it beats with the incoming signal to produce an i-f signal for various frequency components present in the spectrum of the received signal. These signals are presented on a cathode-ray oscilloscope as the spectrum pattern.

A block diagram showing waveforms is shown in figure 18-5. The r-f input is applied to a variable attenuator. The attenuator controls the spectrum amplitude, and consists of two carbonized resistance cards inserted in the r-f input end of the waveguide (fig. 18-7). Varying the contour of these cards varies the degree of attenuation. Maximum attenuation is obtained when the cards are flexed together in the center of the waveguides. The attenuator controls the amplitude of the r-f signal either entering the analyzer, or leaving the analyzer when it is used as a signal generator.

The analyzer r-f oscillator is a reflex velocity-modulated klystron. It is frequency modulated by a portion of the sweep voltage, which is applied to the oscilloscope deflection plates. The sweep voltage creates a maximum frequency swing in the oscillator output of 40 to 50 mc. The analyzer r-f oscillator output is matched to the load by the fixed attenuator in the waveguide.

The incoming signal and the output of the r-f oscillator are mixed at the crystal mixer. The output is applied to the input circuit of the tuned i-f amplifier section to produce the intermediate frequency of 22.5 mc.

An absorption-type frequency meter is mounted in the waveguide at a point between the crystal and the r-f oscillator. The meter consists of a cavity with dimensions that can be varied by a front panel control. Each time the frequency of the r-f oscillator passes through the resonant frequency of the frequency meter, the meter absorbs some of the r-f power. This causes a sharp decrease in the amplitude of the output of the r-f oscillator that is applied to the crystal. The change in amplitude is caused to appear on the scope as a sharp pip, which is used as a frequency marker. The pip occurs when the frequency of the r-f input coincides with the frequency of the meter.

The output of the crystal mixer is applied to the input circuit of the 22.5-mc intermediate frequency amplifier. The output of the i-f amplifier section is applied to the oscillator-converter.

The oscillator-converter contains a local oscillator operating at a frequency of 19.5 mc.

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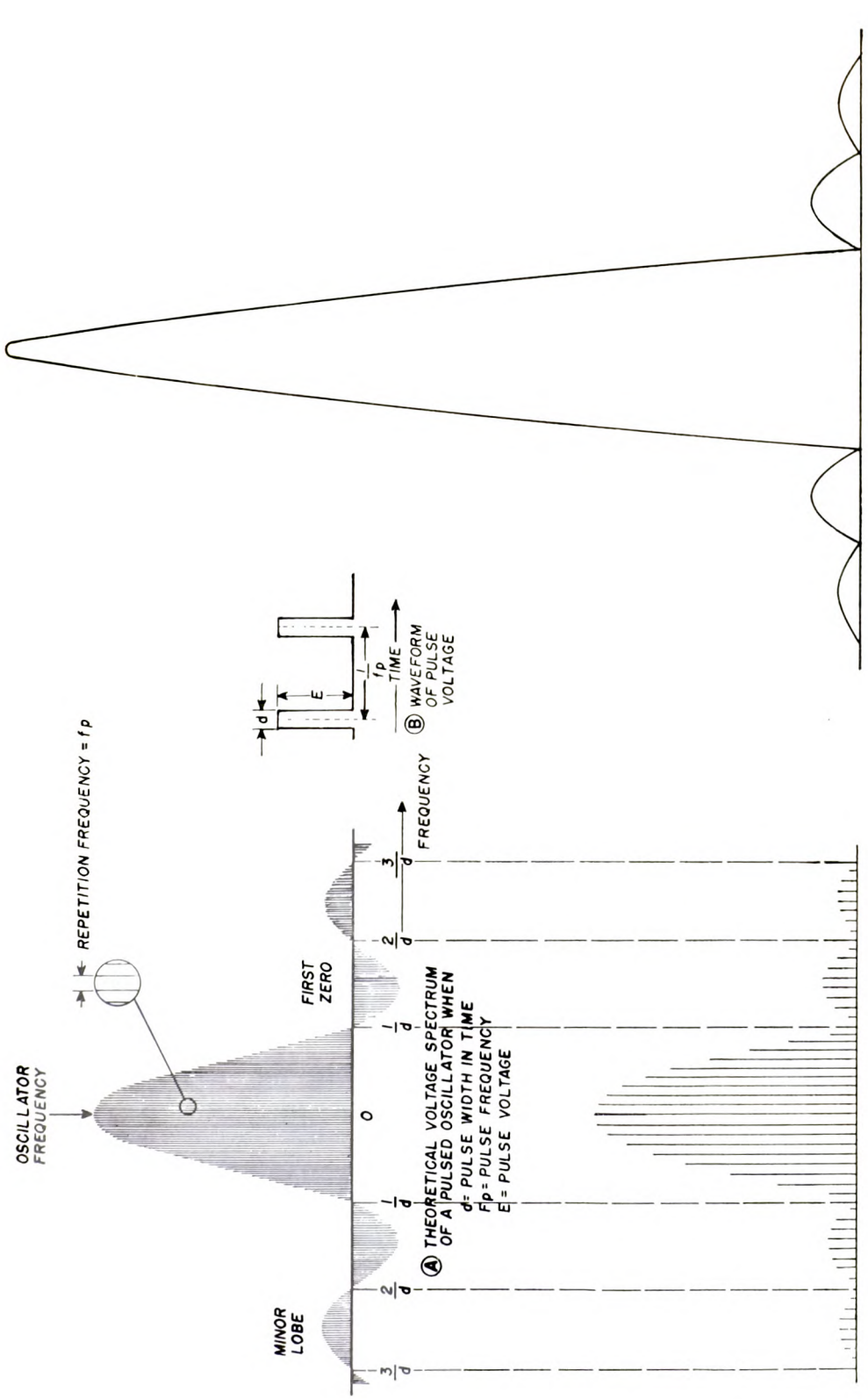
OSCILLATOR FREQUENCY
REPETITION FREQUENCY = fp
MINOR LOBE
FIRST ZERO
FREQUENCY
Ⓐ THEORETICAL VOLTAGE SPECTRUM OF A PULSED OSCILLATOR WHEN
d= PULSE WIDTH IN TIME
Fp= PULSE FREQUENCY
E= PULSE VOLTAGE
Ⓑ WAVEFORM OF PULSE VOLTAGE
TIME
Ⓒ EFFECTIVE VOLTAGES DERIVED FROM ABOVE SPECTRUM
Ⓓ ACTUAL POWER (E²) SPECTRUM ENVELOPE OF SPECTRUM Ⓐ (PLOTTED TO DIFFERENT SCALE) AS SEEN ON TS-148/U

Figure 18-4.—Relationship between a voltage spectrum and a power spectrum.

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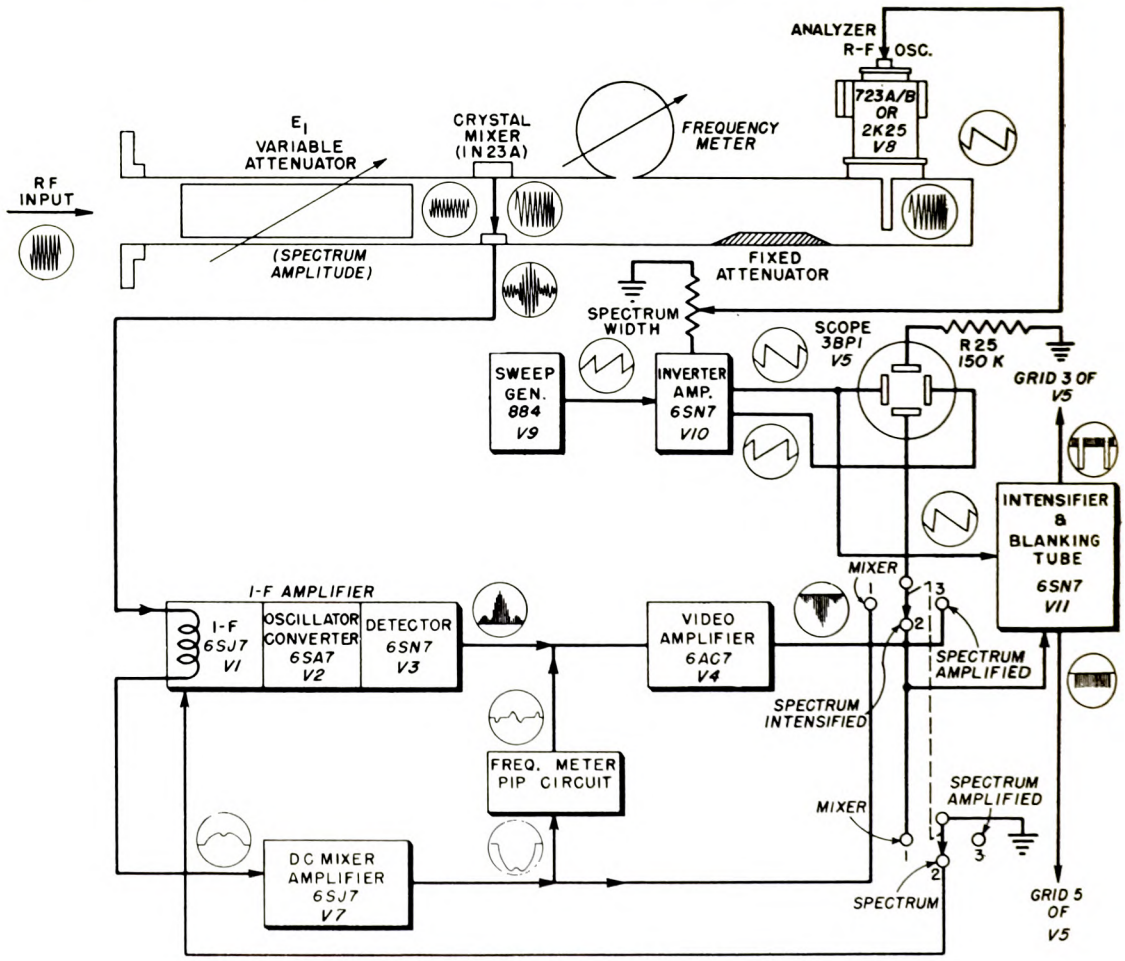

Figure 18-5.—Spectrum analyzer block diagram.

The intermediate frequency is changed to three megacycles in this stage by beating the 19.5-mc frequency against the 22.5 mc from the crystal mixer. The sharply tuned plate circuit of the oscillator-converter produces a 50-kc circuit bandpass. The output is applied to the detector stage.

The detector section functions as an infinite impedance detector. The rectified output of the detector is applied to the video amplifier.

The video amplifier stage has two inputs. One is the output of the detector, and the other the frequency meter pip obtained from the d-c amplifier through a differentiating circuit. The output of the video amplifier is coupled to the vertical deflecting plates of the oscilloscope. A portion of the video output is used to trigger the intensifying stage.

When the selector switch on the front panel is in the MIXER position (position 1), the d-c mixer amplifies the output of the crystal mixer and applies it through the selector switch to the vertical plates of the cathode-ray tube for test purposes. In the other two positions of the selector switch, the output of the d-c mixer is applied to the video amplifier. The end pips of the waveshape appearing at the video amplifier grid are derived by differentiation in the plate circuit of the d-c mixer amplifier. The center pip corresponds to the point of resonance of the frequency meter.

The sweep oscillator is a gas tube relaxation oscillator that generates a sawtooth output at frequencies between 10 and 30 cps. The output of the sweep oscillator is used to drive the sweep inverter amplifier.

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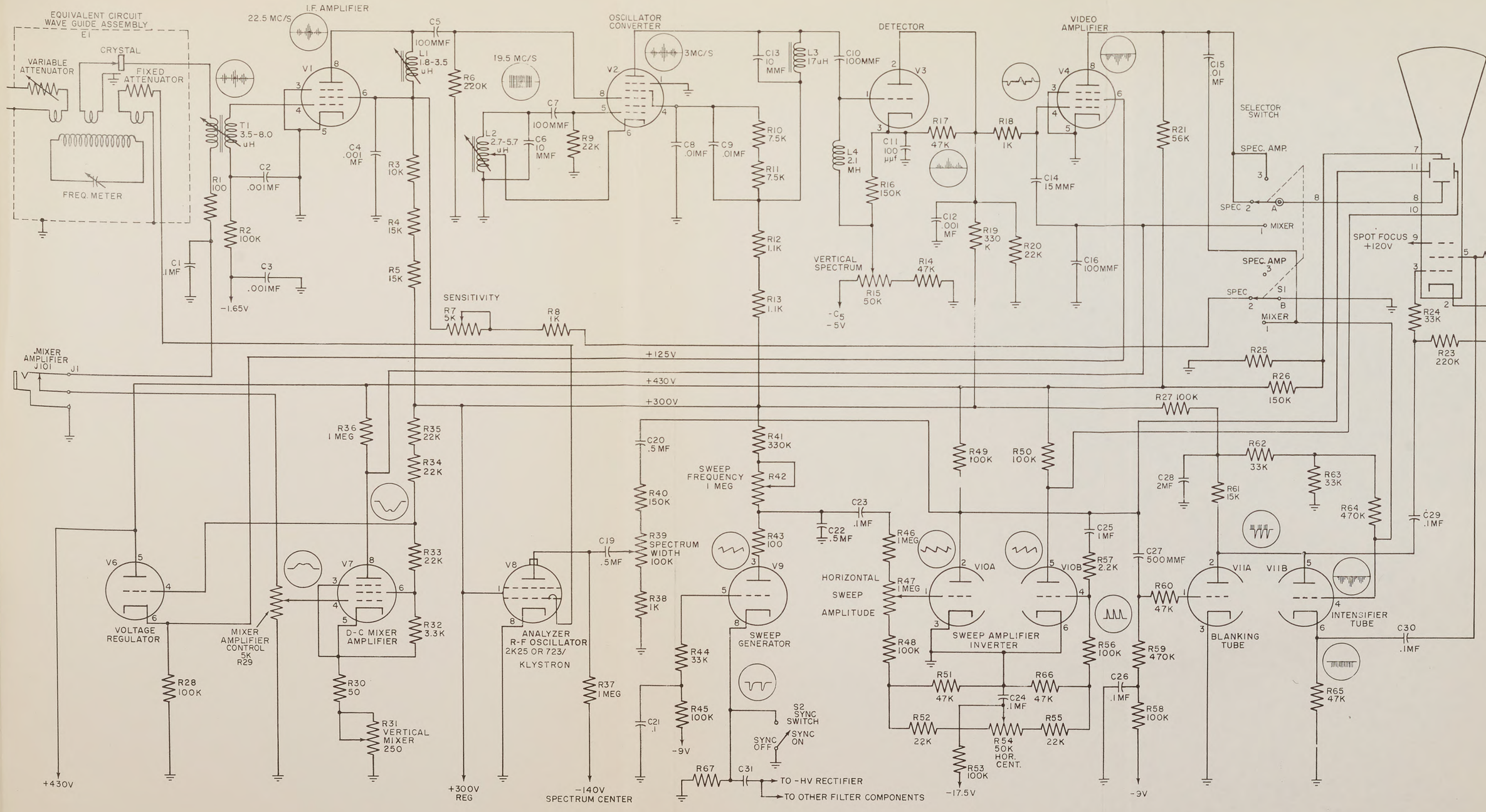

Figure 18-6.—Spectrum analyzer schematic diagram.

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The sweep inverter amplifier provides push-pull deflection voltages for the horizontal plates of the cathode-ray tube. It also supplies modulating voltage to the analyzer r-f oscillator and trigger voltage for the blanking section.

When the selector switch is placed in either of the SPECTRUM positions (positions 2 or 3 on the front panel), the video amplifier signal triggers the intensifier tube. The output thus produced is coupled to the intensifying grid of the cathode-ray tube.

The input voltage from a section of the sweep inverter amplifier is used to trigger the blanking tube. The output of this tube is applied as blanking voltage to eliminate the return trace that would otherwise appear on the scope screen.

CIRCUIT OPERATION

Attenuator

A schematic diagram of the spectrum analyzer is shown in figure 18-6. R-f signals from an outside source enter the waveguide and pass through the variable attenuator to the crystal mixer. The attenuator (fig. 18-7) controls the amount of signal voltage that reaches the mixer, and therefore controls the amplitude of the pattern on the screen. The control knob for the variable attenuator is marked, SPECTRUM AMPLITUDE, on the front panel (fig. 18-1).

The attenuator consists of two carbonized coated resistance strips, which lie flat against the walls of the waveguide. The ends are fastened down but still allow the strips to slide slightly as they are bowed into the center of the guide. As the strips move inward, they intersect more and more of the r-f energy and thus cause the attenuation to increase. The range of the attenuator is from 70 db to 3 db.

Analyzer R-F Oscillator

The analyzer r-f oscillator, V8, uses a Type 2K25 or 723A/B Shepard-Pierce tube. A sawtooth voltage from the sweep inverter amplifier (discussed later) is applied across the spectrum width control, R39. This voltage is coupled through C19 to the repeller electrode of the klystron, V8.

V8 forms a reflex-type, variable-cavity oscillator. The tube contains an electron gun to emit electrons into the cavity. The cavity is maintained at 300 volts positive with respect to its cathode, which is grounded. The +300 volts is obtained from a regulated supply. The repeller electrode is connected to a -140-volt source through R37.

The electrons from the cathode are accelerated as they pass through the positive grids. The a-c potential of the grids (caused by the movement of electrons in and around the grids) changes at the frequency of the oscillator. If a group of electrons arrive at the first grid when it is positive with respect to the second, the electrons are accelerated and pulled away from the electrons that immediately follow. If the second grid is still negative when the group of electrons reach it, the group will be slowed and bunched closer together.

If the second grid is not negative, the electrons pass through it to the negative repeller where they are repelled back into the cavity. If the electrons arrive back at the cavity in phase with the a-c potential of the cavity, they reinforce the bunching effect, giving up energy to the cavity.

The result of this action is a velocity-modulated stream of electrons flowing in the cavity. As the bunched electrons pass through each grid, it becomes alternately positive and negative. The frequency of this oscillating action is determined by the spacing between the grids, the volume and shape of the cavity, the space between the repeller and the cavity, and the voltage on the repeller. Assuming that the d-c potentials on the repeller and the cavity are constant, the voltage and spacing of the repeller act to control the transit time of the electrons, which fixes the frequency of the oscillation.

The analyzer r-f oscillator may be tuned in either of two ways. One way is by means of a mechanical screw (fig. 18-8). The screw, operated by the oscillator frequency control on the front panel, places a stress on a pair of struts on the side of the tube. As the struts are lengthened they expand the cavity to increase the spacing between the two grids and vary the capacity in the resonant cavity.

The second method used to tune the analyzer r-f oscillator involves varying the voltage on the repeller electrode. The spectrum center knob on the front panel is a fine frequency control, which varies the d-c potential on the repeller. The spectrum center potentiometer is not shown on the schematic. Actually, the klystron will only operate within definite ranges of repeller voltage. The ranges are called modes. The analyzer r-f oscillator normally

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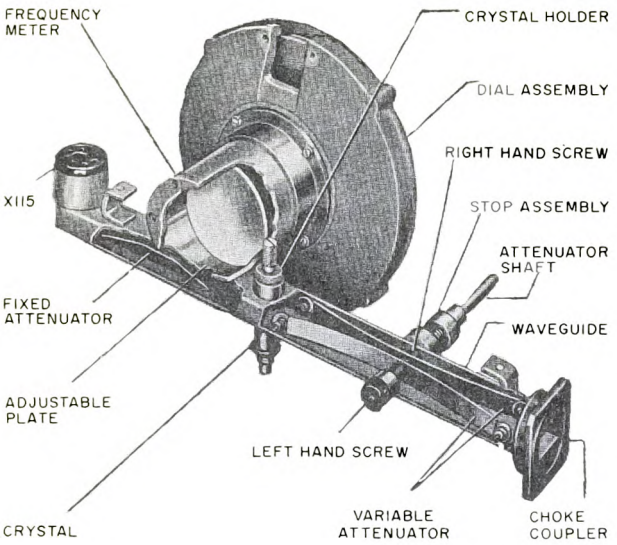

Figure 18-7.—Waveguide assembly, cut-away view.

has a d-c potential of -140 volts applied to the repeller electrode, and makes use of the modes that appear in the -100 to -180-volt range.

It is necessary to frequency modulate the analyzer r-f oscillator V8 in order to display on the scope screen each frequency component of the signal under test. The pulse coupled into the waveguide through the variable attenuator contains many frequencies. The klystron frequency is caused to vary by sweep modulation. The two varying frequencies, when heterodyned at the crystal mixer, will produce a 22.5-mc output at various times or at various points along the frequency sweep of the r-f oscillator. Figure 18-9 shows the spectrum, and how the two frequencies are heterodyned to produce the spectrum components.

The signal at the beginning of the sweep represents the lowest frequencies in the lower sideband (fundamental frequency minus the higher order harmonics) and is followed in turn

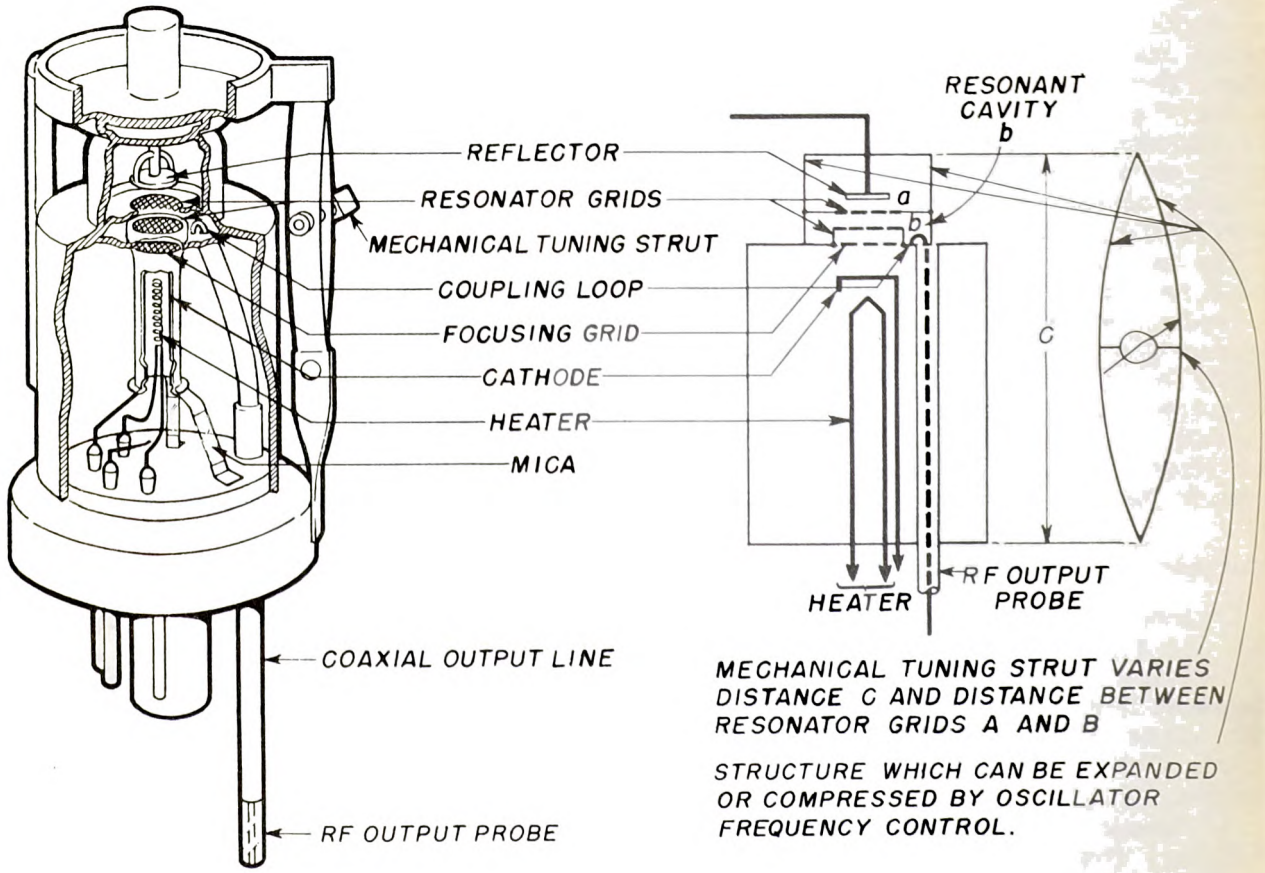

Figure 18-8.—Analyzer r-f oscillator tube.

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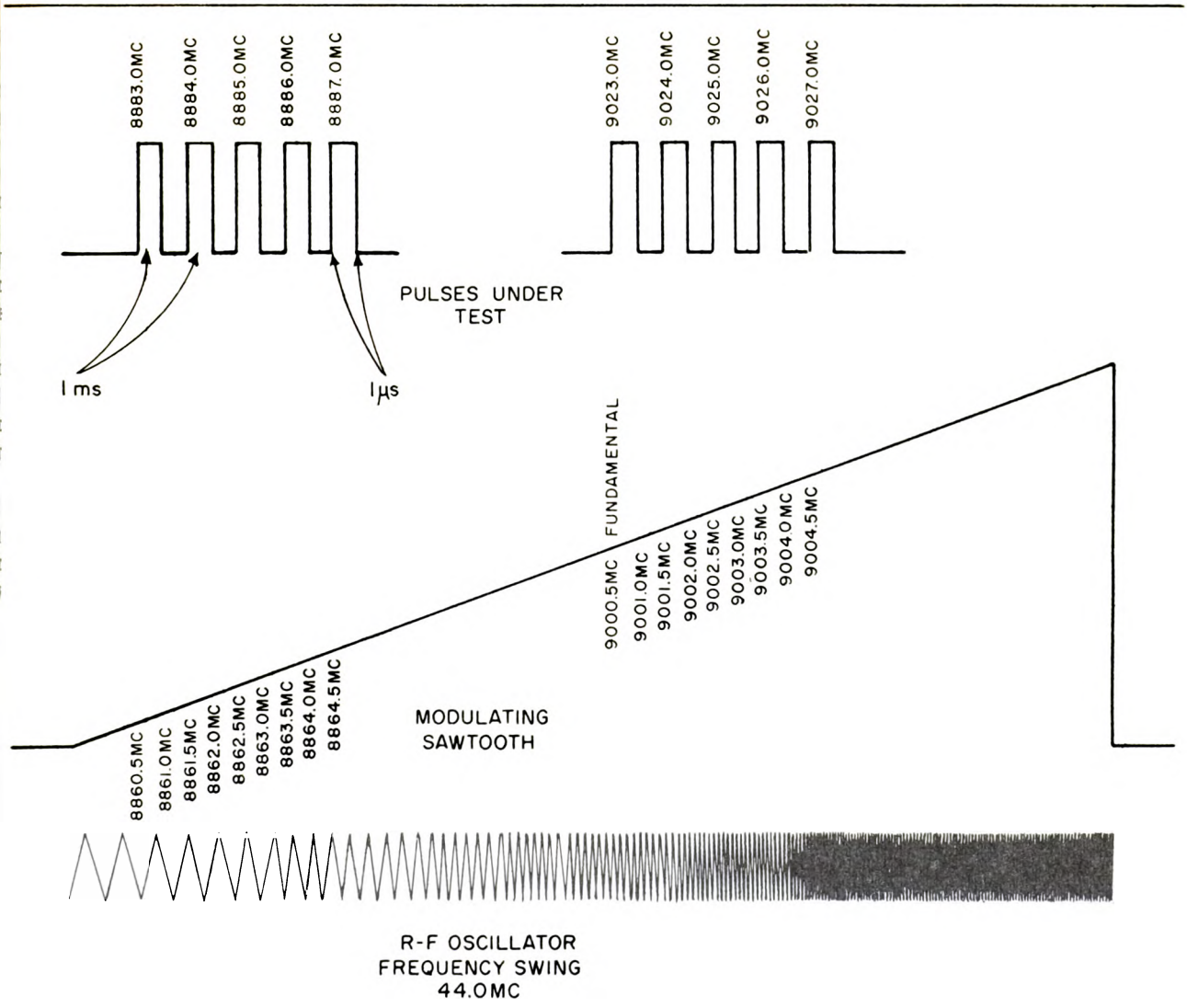

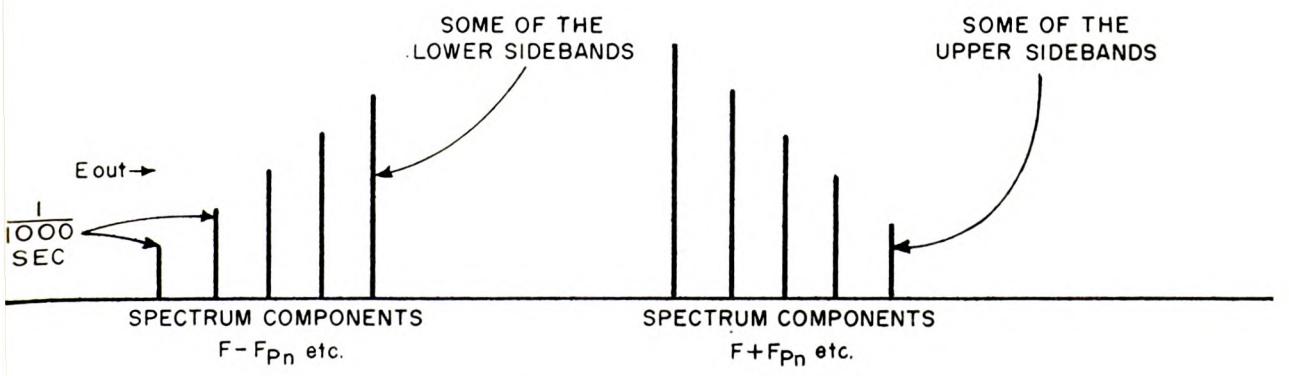

Figure 18-9.—Method of producing spectrum frequencies.

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by the higher frequencies in the lower sideband (fundamental frequency minus the higher order harmonics) until the fundamental frequency is reached. The succession of i-f signals continues through the upper sidebands until the highest frequency is reached. Since the trace on the scope is synchronized in time with the frequency swing of the analyzer r-f oscillator (as discussed later), the sideband frequencies will be spread out from left to right as shown in figure 18-9.

The output of the analyzer r-f oscillator is passed from the coupling loop inside the cavity of the klystron, V8 (fig. 18-6), through a fixed attenuator, to the crystal mixer. The fixed attenuator isolates the analyzer r-f oscillator from the rest of the waveguide so that it is unnecessary to make any adjustments on the mixer when tuning over the frequency range. A portion of the oscillator output is used by the crystal in the heterodyning process. The portion not used in this process moves past the variable attenuator and out into the waveguide. This portion of the r-f is of particular importance when the analyzer is used as a signal generator.

Frequency Meter

The power from the r-f oscillator goes past the frequency meter (fig. 18-6) where a small portion of the energy is absorbed by the cavity of the meter. The frequency meter is a high Q cylindrical cavity-type meter with a mechanically variable length. The frequency range of the meter is 8470 to 9630 mc.

Each time the frequency of the analyzer r-f oscillator sweeps through the frequency to which the frequency meter is tuned, the meter absorbs some of the r-f power. The absorption results in a sharp reduction of the analyzer r-f oscillator component of the crystal current. This reduction, or pip, is applied to the d-c mixer amplifier and is used to indicate the coincidence of the frequency meter with unknown frequencies on the scope screen. The reduction of the amplitude of the analyzer r-f oscillator frequency causes a corresponding reduction in the amplitude of the i-f signal.

Crystal Mixer

The crystal mixer (fig. 18-6) is used to rectify the heterodyne frequency produced by mixing the f-m output of the analyzer r-f oscillator with the r-f input signal in the waveguide. When the input signal is 22.5 mc above or below the frequency of the analyzer r-f oscillator, the i-f amplifier responds to the output of the crystal.

The crystal is located between the frequency meter and the variable attenuator (fig. 18-7). The current through the crystal represents the rectified components of the signals. The output of the crystal is coupled by T1 (fig. 18-6) to the input of the i-f amplifier. The signal contains a rectified component that is proportional to the power output of the analyzer r-f oscillator with respect to the frequency.

I-F Amplifier

The i-f amplifier, V1, is a conventional amplifier using an iron-core, tuned-coil secondary (by distributed capacitance) and a fixed d-c bias of -1.65 volts at the grid. The stage amplifies the i-f signal and prevents the output of the oscillator-converter, V2, from appearing in the waveguide.

The output of the analyzer r-f oscillator varies at an audio rate because of the modulation by the 10- to 30-cycle sawtooth voltage, as explained earlier. The selective characteristics of T1 prevents the audio component of the oscillator from entering the i-f amplifier.

V1 has a tuned inductive plate load, which produces a 500-kc bandpass. The output of V1 is a series of pulses of 22.5 mc. Each pulse represents a section of the spectrum. The output of V1 is coupled by C5 to the injector grid of the oscillator-converter, V2.

The spectrum position of the selector switch, S1, inserts potentiometer R7 and series resistor R8 between the screen of V1 and ground. The resulting increase in current through R3, R4, and R5 causes a drop in the screen and plate voltage from 110 to approximately 40 volts. This causes a 25-db reduction in the gain of the i-f amplifier.

Oscillator-Converter

The primary purpose of the oscillator-converter is to provide narrow bandwidth characteristics for the analyzer. This is accomplished by a 19.5-mc Hartley oscillator consisting of tank circuit L2 and C6, and a triode oscillator tube consisting of pins 4, 5, and 6 of V2. The 19.5-mc signal combines with the 22.5-mc incoming i-f signal to produce a

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3-mc difference frequency. The bandwidth produced in this stage is 50 kc at the half-power points.

C7 and R9 provide grid leak bias for V2. The oscillator voltage on grid 5 causes pulses of tube current to flow during the positive peak of each oscillator cycle. Grid 8 receives the 22.5-mc i-f signal from V1. The pulses of tube current arriving at grid 8 form a virtual cathode whose emission is controlled by the signal voltage. Since the virtual cathode does not exist except on the peaks of the oscillator cycles, the plate current varies at the heterodyne frequencies of the original intermediate frequency and the oscillator frequency.

The signal is amplified in V2 and developed in the plate circuit across the tuned high Q inductor, L3.

Detector

The i-f signal is coupled to the detector, V3, by C10. V3 operates as an infinite impedance detector to minimize the loading effect on L3. The operating point of V3 is determined by the setting of the vertical spectrum control, R15. This control has no effect on V3 other than to change its cathode-to-plate potential slightly. The main purpose of the vertical spectrum control is to regulate the bias on the video amplifier, V4 which is discussed later. The fixed bias voltage for V3 is developed from the R14, R15 junction to ground.

The voltage across the 150-kilohm resistor, R16, maintains a cathode bias on V3 sufficient to hold the tube very near plate current cutoff. Any a-c signal appearing at the grid is rectified or detected since the negative half cycle cuts off V3 while the positive half cycles control the amount of plate current flow.

The cathode of V3 is connected to the control grid of V4 through coupling resistors R17 and R18. A low value of plate voltage for V3 is obtained from the junction of R19 and R20, which forms a voltage divider from the +300-volt supply to ground. C12 places the plate at r-f ground potential.

Video Amplifier

With the exception of the direct coupling between the video amplifier grid and the detector cathode, V4, forms a conventional amplifier circuit. When the selector switch, S1, is in the SPECTRUM of SPECTRUM AMPLIFIED positions, any change in the plate or grid voltage of V4 produces a deflection on the vertical plates of the oscilloscope. Likewise, vertical centering is accomplished by adjusting the d-c bias on the grid of V4. The bias is obtained from the arm of the vertical spectrum control, R15, through R16, R17, and R18 to the control grid of V4. The voltage across R15 develops a potential high enough to keep the scope pattern centered for variations in the a-c power.

Screen voltage for the video amplifier is obtained from a voltage regulator, V6. The control grid of V6 is connected through R34 and R35 to the regulated +300-volt supply. Plate voltage is obtained from the +430-volt B supply. The resulting cathode voltage of +125 volts to ground (caused by the tube conduction through R28) is applied to the video amplifier screen.

The d-c mixer amplifier, V7, supplies an output, as shown in the figure. The method by which this output is developed is discussed later.

The output of the d-c mixer amplifier is applied through C14 to the control grid of V4, R14 through R18, and C14 form a differentiating circuit at the V4 grid. Remember that the d-c mixer amplifier signal was derived from the modulating signal at the analyzer r-f oscillator. The sharp pip in the d-c mixer amplifier output originated from the effects of the frequency meter absorption at its resonant frequency. A harmonic content in the V7 output is evident by its nonsinusoidal appearance.

C14 and its associated resistance pass the high frequency components at the beginning of the curve, the frequency meter pip, and the end of the curve. This results in a waveform at the V4 grid shown in the figure. The two pips on either side of the frequency meter pip indicate the limits of the sweep of the analyzer r-f oscillator. This, in turn, corresponds to the limits of the observed spectrum.

The output of V4 is negative and is applied to the lower vertical plate of the cathode-ray tube when the selector switch, S1, is in the SPECTRUM or SPECTRUM AMPLIFIED position. C15 couples a portion of the V4 output to the intensifier tube, V11B, when S1 is in either of the above positions.

Sweep Generator

The sweep generator, V9, employs a gas-filled triode tube in a relaxation-type sawtooth

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generation circuit. The stage generates a voltage to control both the scope trace and the frequency sweep of the analyzer r-f oscillator.

A fixed bias obtained from the negative voltage supply (not shown) is applied through R44 and R45 to the grid of V9. A small synchronizing voltage from the negative high-voltage filter circuit may be applied to the cathode of V9. This voltage is obtained from the junction of R67 and C31 when the sync switch, S2, is in the SYNC-ON position. In the SYNC-OFF position, S2 shorts the cathode resistor, R67. The ripple voltage across C31 also appears across R67. The frequency of the line voltage is constant, which makes it possible to synchronize the sawtooth modulation to the analyzer r-f oscillator with the pulse modulation of any radar equipment that is synchronized to the line frequency.

V9 is normally cut off. With the sync switch, S2, in the SYNC-OFF position, the plate voltage of V9 rises above the firing potential, which is fixed by the bias on the grid, and V9 conducts. C22, which charges to the firing potential, can now discharge through the conducting tube. When C22 sufficiently discharges, the plate voltage of V9 drops to the extinction point of the tube, and conduction ceases. C22 now begins to charge through R42 and R41 to the B supply voltage. R42 serves as the sweep frequency control since it varies the charge time of C22. When C22 charges sufficiently to raise the plate voltage of V9 to a point where it overcomes the fixed bias on the grid, the tube again acts as a closed switch across C22, and the action recurs. The resulting sawtooth output controls the sweep of the cathode-ray tube and serves as modulating voltage for the analyzer r-f oscillator. Since the part of the sawtooth used is taken from the first portion of the capacitor charge, its rise time is effectively linear with respect to time. The output of V9 is coupled by C23 to the control grid of the sweep inverter amplifier, V10.

A 60-cps synchronizing voltage is provided at the V9 cathode from the C31-R67 junction when the sync switch, S2, is in the ON position. C31 and R67 form a voltage divider in the high voltage filter circuit (not shown). The negative pulses at the V9 cathode synchronize the sweep generator to a submultiple of the line frequency.

Sweep Inverter Amplifier

The grids of V10 are connected across a balanced bridge circuit. The bridge consists of V10A, the resistance from the arm of R47 through R48 and R51 in one half, and V10B, R56, and R66 in the other half. R52, R54, and R55 are connected across R51 and R66 such that they parallel a section of each half of the bridge. The arm of R54, the horizontal centering control, is connected to -17.5v and is adjusted to balance the d-c output of each section of the bridge. Because the plates of V10 are directly connected to the horizontal deflection plates of the cathode-ray tube, the position of the trace depends upon the d-c potentials on the plates of the tube.

The amplitude of the input signal at V10A from the sawtooth generator, V9, is controlled by the adjustment of the horizontal sweep amplitude control, R47. This control regulates the amplitude of the horizontal sweep across the cathode-ray tube.

A portion of the output of V10A appears across C25, R57, R56, and R66. This circuit comprises a voltage divider to obtain the proper amount of attenuation in the signal at the grid of V10B to insure a balanced output from plates 2 and 5. A portion of the plate 2 output is also coupled through C27 to the control grid of the blanking tube, V11A (discussed later). Another portion of the plate 2 output is coupled by C20 through R40, R39, and R38 to ground. That part of the sawtooth voltage developed across R39 (from the arm to ground) is used as modulating voltage for the analyzer r-f oscillator as discussed earlier.

Blanking Amplifier

If the back trace (or sweep return) should be visible on the screen, another spectrum would appear with sufficient intensity to blend into the desired pattern and render it practically useless. The return of the trace is made invisible by the action of V11A.

Fixed negative bias of -9v for V11A is obtained from the negative voltage supply, which is sufficient to hold the tube cutoff. The plates of V11 are tied together because both outputs are applied to the control grid of the cathode-ray tube, V5.

C27 and R59 form a differentiating circuit, which removes the low-frequency components of the sweep voltage to produce short positive pulses of voltage. These pulses are applied to the grid of V11A through grid limiting resistor, R60. The output at the V11A plate consists of negative pulses synchronized with the decay time of the sawtooth sweep voltage. The pulses are coupled by C29 to the control grid of the cathode-ray tube to cut off the electron beam for the

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duration of the retrace. The two plates of V11 are not active simultaneously. The blanking signal occurs during the decay time of the sweep while the intensified spectrum pattern appears during the linear portion of the sweep voltage. The development of the intensifying pulses is explained in the next few paragraphs.

Intensifier

The intensifier stage, V11B, is provided to intensify or brighten the individual signal pulses that make up the spectrum on the screen. This provides better pattern definition and makes it unnecessary to turn up the manual intensity control to an abnormal level of brightness in order to be able to distinguish the dim spectrum pulses.

When the selector switch, S1, is placed in the SPECTRUM or SPEC. AMP. position, the negative output of the video amplifier, V4, is coupled through C15 and developed by R64 and R63 at the control grid of V11B. The plate of V11B therefore develops a positive pulse each time the video amplifier delivers a pulse of voltage to the vertical plates of V5. The pulse from the video amplifier is negative while the signal from V11B, which appears at the control grid of V5, is positive. The intensifying pulses from V11B increase the number of electrons in the beam and brighten the pattern (not the sweep) during the time each signal pulse is delivered to the vertical plate of V5.

The negative pulses appearing at the cathode of V11B are applied through C30 to the focusing anode of the scope to provide an instantaneous focus correction while the grid is being intensified. The d-c grid to ground voltage is approximately +40 volts developed by R63, and the voltage across R65 is approximately +43 volts. This results in a negative 3 volts bias for the tube. The plate potential is obtained from the junction of R62 and R63, and is approximately +70 volts. When the negative input signal at the grid of V11B exceeds 45 volts, the grid of the tube is driven past cutoff. The pulses from the cathode and plate of V11B are limited to about -40 volts and +12 volts, respectively.

When the selector switch, S1, is in the MIXER position, the grid of V11B is grounded to prevent both the i-f signals and the frequency meter pip from being intensified.

D-C Mixer Amplifier

The d-c mixer amplifier is a conventional amplifier used to produce a test pattern of the output of the analyzer r-f oscillator and to produce the frequency meter pip on the spectrum. The i-f component of the crystal mixer output is coupled by T1 to the control grid of V1. The low-frequency component, associated with the modulating frequency and containing the frequency meter pip, is fed through R1 and R29 to ground. That portion of the voltage appearing at the arm of the mixer amplifier control, R29, is applied to the grid of the d-c mixer amplifier, V7. C1 bypasses the i-f frequency to ground.

Screen and cathode voltages for V7 are obtained from the +300-volt regulated supply. Vertical mixer control R31 adjusts the bias level of V7 and therefore the d-c plate potential. Since the plate of V7 is connected directly to the vertical plates of the cathode-ray tube when S1 is in the MIXER position, R31 is considered the vertical centering control for this position of the switch.

C16, connected between the plate of V7 and ground, places the plate at r-f ground potential, preventing the i-f component (which may still be present) from appearing as a part of the signal at the vertical deflection plates.

The output of V7 is a slowly changing voltage that follows the envelope of the output of the analyzer r-f oscillator. The steep sections of the output wave are differentiated by C14 and the grid resistance of V4 to produce the frequency meter pip on the spectrum, as discussed earlier.

Phone jack J1 removes the output of the analyzer crystal mixer and substitutes the AFC radar crystal mixer when connected to the jack through the proper cable.

PROCEDURE FOR OBSERVATION OF SPECTRA

Magnetron or Other Oscillator Spectra

Assuming that the spectrum analyzer has been installed and turned on, the selector switch (fig. 18-1) should be placed in the MIXER position. In this position of the switch, adjustments may be made on the analyzer r-f oscillator. The spectrum width control should be turned in a clockwise direction until its setting is nearly maximum.

After all necessary adjustments have been made on the analyzer, and the desired width of the scope trace is set, place the selector switch in the SPECTRUM or SPECTRUM AMPLIFIED position. Use the spectrum amplified position if

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there is no likelihood of interference from any adjacent radar equipment. If interference occurs, spectrum position should be used.

Point the waveguide, which is accessible through the side door shown in figure 18-1, at the radar antenna or to the magnetron output. Coupling the waveguide too close to the source will burn out the attenuator and crystal mixer in the analyzer. Likewise, if the coupling is too close, the probability of magnetron pulling is increased.

An antenna horn and fittings (component parts of the TS-148/UP) may be used if the former method does not supply satisfactory results (fig. 18-10). This method can be used to pick up the output of the magnetron at distances up to 100 ft if the two antennas are pointing directly at each other.

The operator should never attempt to increase the amplitude of the spectrum trace by increasing the coupling until it is ascertained that the amplitude cannot be increased by adjusting the spectrum amplitude control (attenuator) or by placing the selector switch in its SPECTRUM AMPLIFIED position, or both. To do so would burn up the resistance cards in the attenuator and burn out the crystal, even though the equipment is turned off.

The antenna horn should be pointed directly to the source under test. In the case of an oscillator, it is seldom necessary to remove the oscillator cover. The magnetron and local oscillator spectrums may be observed simultaneously if a proper location for the antenna horn is found that produces sufficient attenuation to the output of the magnetron.

If a very stable method of coupling is desired, remove the antenna horn and connect the cable to the directional coupler or wave selector (fig. 18-11), which is inserted at some convenient place in the transmission line of the radar equipment. The spectrum analyzer will then receive approximately one percent of the power in the directional coupler. This method is very satisfactory when it is necessary to check the antenna while it is in motion to see if it is pulling the frequency of the magnetron.

Observation of Patterns

Adjust the spectrum amplitude control to obtain the desired amplitude on the screen. Rotate

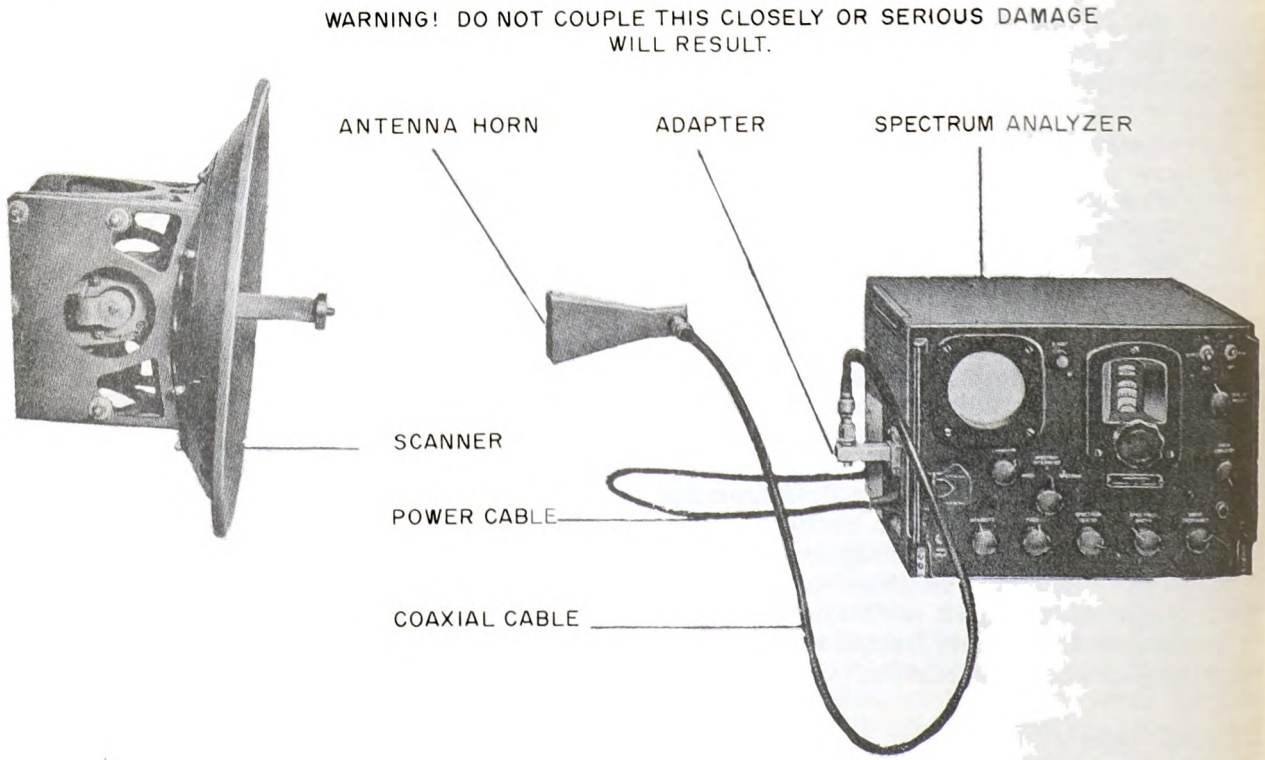

Figure 18-10.—Use of antenna horn.

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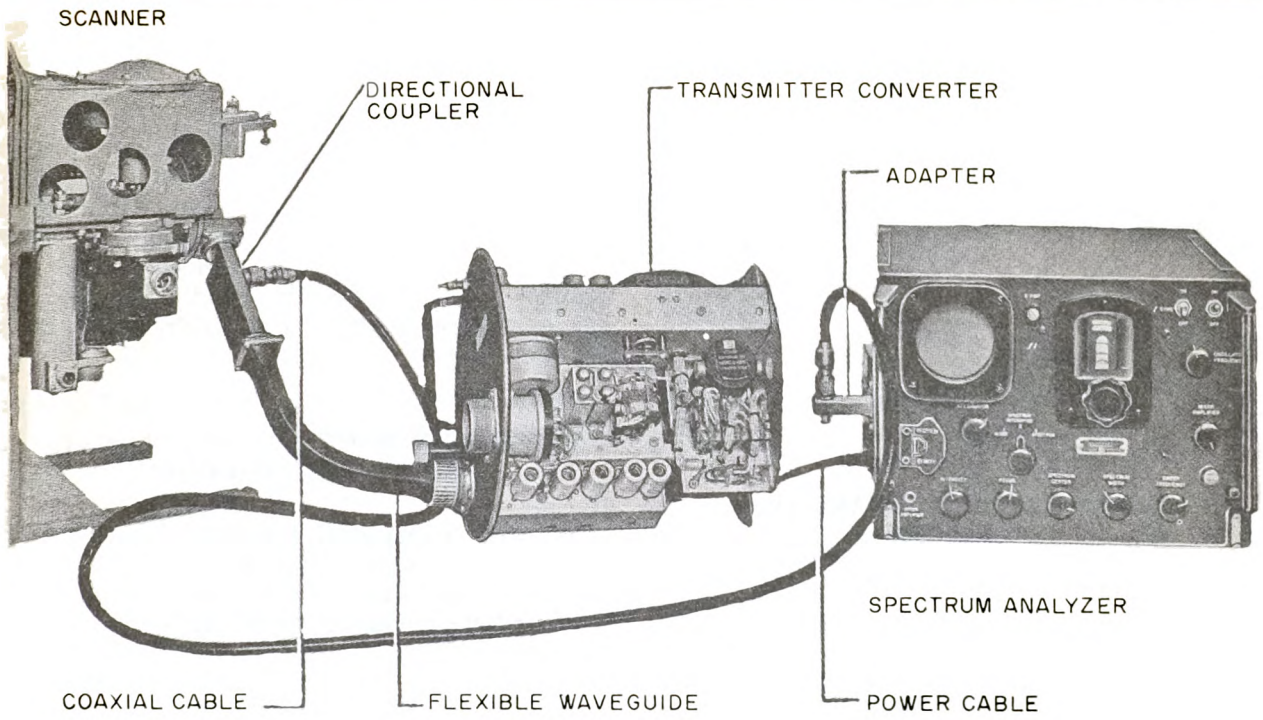


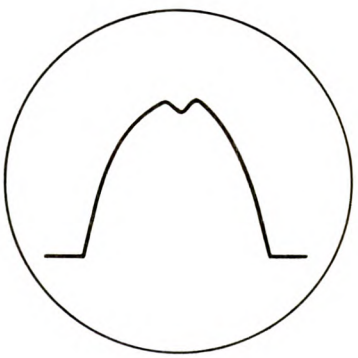
Figure 18-11.—Use of the directional coupler

the spectrum width control counterclockwise until the spectrum gains sufficient detail to make observation easy. If more pulses are desired in the spectrum, turn the sweep frequency control counterclockwise to a lower sweep frequency.

FREQUENCY MEASUREMENT

The frequency meter pip that is visible when the selector switch is in the MIXER position (fig. 18-12) represents the actual frequency of the frequency meter, as read on the frequency meter dial. When the selector switch is in either of its other two positions for frequency measurements, the reading on this dial does not represent the frequency being measured. The unknown frequency must be computed from the dial reading by adding or subtracting the 22.5-mc intermediate frequency of the analyzer.

Whether the intermediate frequency is added or subtracted depends on the image the frequency meter pip is made to coincide. Two images will appear for each component of the spectrum (fig. 18-13). The frequency of the analyzer r-f oscillator increases as the sweep progresses from left to right across the screen. This action causes an image to be produced below and above the 22.5-mc intermediate frequency. In most cases, both images can be made to appear on the screen although, in actual practice, one is



a. MIXER POSITION

NO SIGNAL PRESENT EXCEPT THAT OF ANALYZER R-F OSCILLATOR

Figure 18-12.—Frequency meter pip (selector switch in MIXER position).

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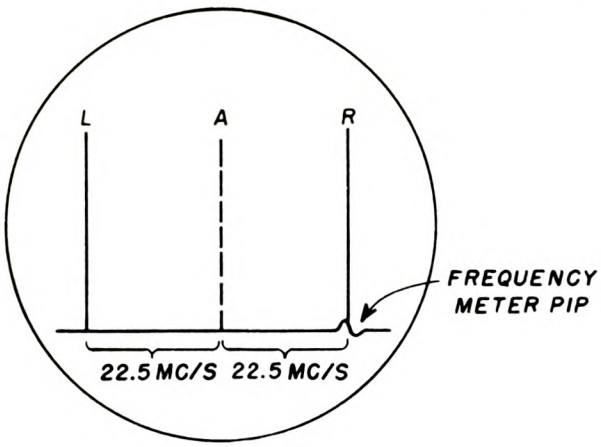

Figure 18-13.—Right and left-hand patterns.

centered on the scre so that the other does not appear.

If the image selected for frequency determination is the one on the right, it is necessary to subtract 22.5 mc from the dial reading. The remainder thus obtained represents the frequency of the unknown signal. If the left image is selected, it is necessary to add 22.5 mc to the dial reading. Since the two images appear above and below the true frequency by 22.5 mc, it is obvious that the two reading may be made directly from the meter, and the average taken to obtain the unknown frequency.

Another very accurate method may be used if the unknown signal is stably coupled to the analyzer. The method consists of tuning the frequency meter until its pip is equi-distant from the two images (fig. 18-14). While tuning the meter, both images must be carefully observed. At the instant the resonant frequency of the frequency meter equals the frequency of the unknown signal, it will be possible to see both images dip simultaneously. The reduction in amplitude occurs because the frequency meter absorbs energy from the unknown signal frequency. In this case, the reading on the frequency dial represents the actual frequency of the unknown signal.

MEASUREMENT OF RADAR LOCAL OSCILLATOR FREQUENCY

There are two similar methods for measuring the frequency of a radar local oscillator. The choice of methods depends upon whether the intermediate frequency is 30 mc or 60 mc/s.

Assume that the radar intermediate frequency is 60-mc cycles, and that the local oscillator frequency is above that of the magnetron. It should be noted that two frequencies 60 mc apart cannot be seen on the same scope trace. This is true because the full width of the i-f above and below the analyzer radio frequency is only 45 mc wide (fig. 18-15). In order to see both patterns simultaneously, a unique use is made of the fact that two spectrum images appear for each signal. In practice, the spectrum center control is used to make the right-hand spectrum of the magnetron appear on the left side of the scope. The left-hand pattern of the radar local oscillator should then appear to the

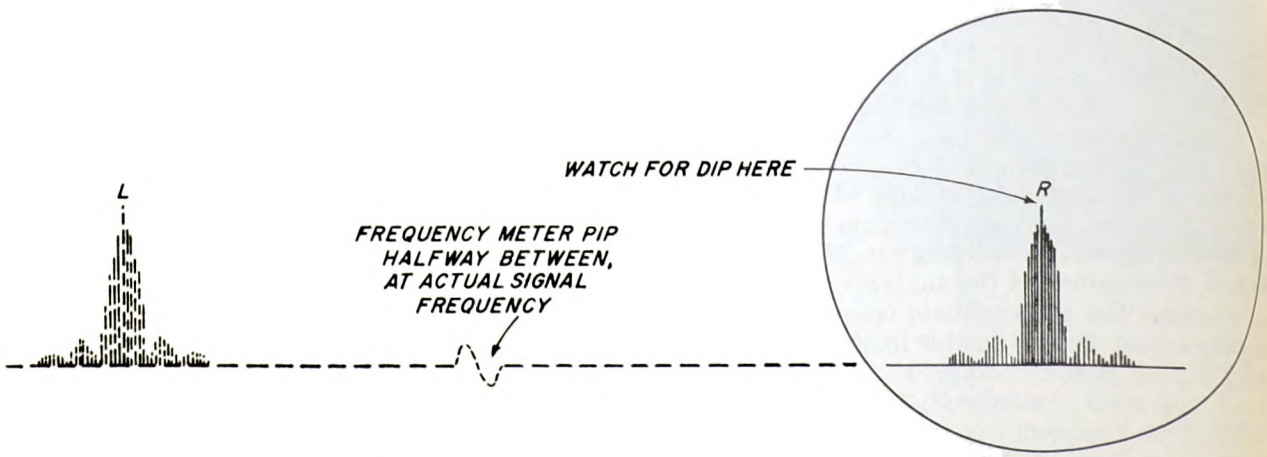

Figure 18-14.—Determination of frequency by tuning frequency meter for dip in spectrum.

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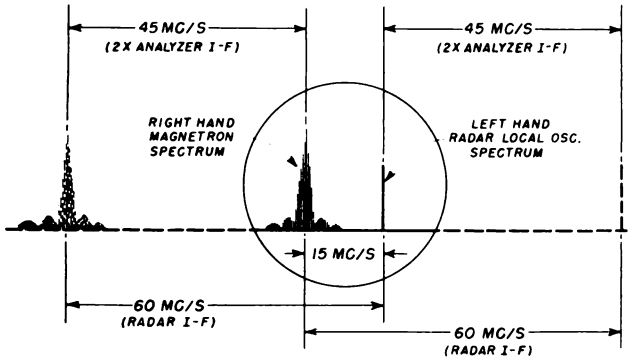

Figure 18-15.—Measuring the frequency of a local oscillator that is 60 mc above the magnetron.

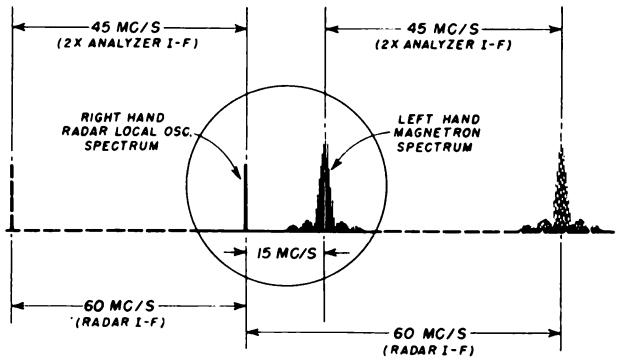

Figure 18-16.—Measuring the frequency of a local oscillator that is 60 mc below the magnetron.

right of this pattern. The pattern represented by the dotted lines in the figure does not appear on the scope, but is shown to illustrate the frequency relationship of the two images that do appear.

The accuracy with which the radar local oscillator tracks the magnetron can be determined by measuring the difference between the frequencies of the two images on the screen. This difference should be 15 mc (using the lower magnetron image as the reference). To determine the actual difference, rotate the frequency dial to set the frequency meter pip under the center of the magnetron spectrum, and note the dial reading. Then set the pip under the radar local oscillator pip, and again note the dial reading. If the difference between the dial readings is 15 mc/s, the radar local oscillator (or AFC discriminator) has been accurately set to follow the excursions of the magnetron frequency.

If the frequency of the radar local oscillator is set 60 mc below the magnetron frequency, the only effect is that the images will be reversed on the screen (fig. 18-16).

If the intermediate frequency is 30 mc/s above, the procedure is similar to that first discussed, except that the position of the images is reversed (fig. 18-17). This is true since by time relationship, the left-hand spectrum of the radar local oscillator is made to appear on the left side of the screen, while the spectrum of the magnetron now appears on the right. The difference in the frequencies is measured just as it was in the foregoing method, and a frequency difference of 15 mc/s indicates that the local oscillator is operating at a frequency 30 mc/s above the magnetron.

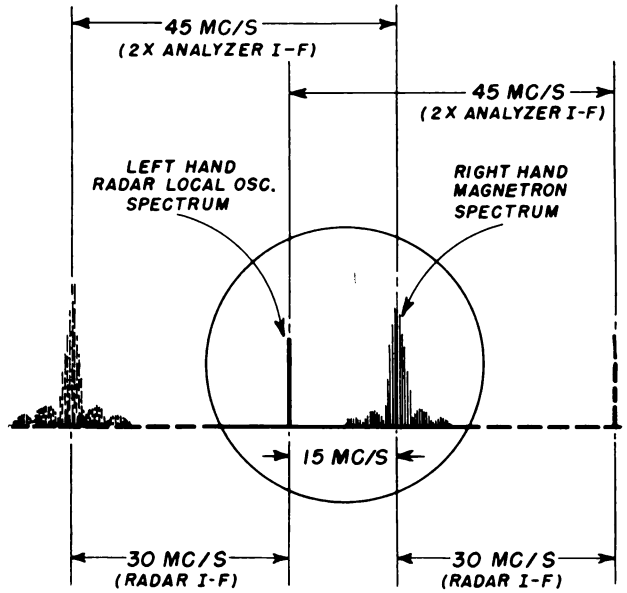

Figure 18-17.—Measuring the frequency of a local oscillator that is 30 mc above the magnetron.

If the radar local oscillator operates below the magnetron frequency, the position of the images will again be reversed. That is, the radar local oscillator spectrum will be on the right and the magnetron spectrum on the left.

SPECTRUM ANALYZER USED AS A SIGNAL GENERATOR

The analyzer may be used as a c-w or f-m signal generator, depending upon whether the

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spectrum width control is set at either of its extreme positions. If it is set completely counterclockwise, the percent of frequency modulation is at a minimum (about 3 mc/s), and the output is essentially a c-w signal. If the control is turned all the way clockwise, the percent of frequency modulation is maximum, and the output has a width of from 40 to 50 mc, depending upon the condition of the tube in the analyzer r-f oscillator.

ABSORPTION WAVEMETER

A wavemeter is a calibrated electric resonator whose resonant frequency can be adjusted to correspond with the frequency of the circuit under test. This is accomplished by two interacting circuits. The wavemeters discussed here (fig. 18-18, A) are designed to absorb some of the energy from the source under test, and are therefore termed absorption wavemeters. Power relationships can also be determined by use of the wavemeter indicator.

The frequency range covered by the meters on this discussion are from 80 to 360 mc, and from 330 to 1220 mc. When the output frequency of a source is to be checked, resonance between the two circuits is indicated on a 0 to 1 miliammeter.

Absorption wavemeters require no external power source. All power is derived from the signal being measured, which is rectified by a crystal installed within the wavemeter circuit (discussed later).

CONNECTIONS

The antenna for the wavemeter plugs into the socket on top of the antenna housing (fig. 18-18, B). One end of the connecting cable is permanently connected to the milliammeter. The free end of cable makes connection to a socket located on the side of the antenna housing. When these connections are made, the wavemeter is ready for use.

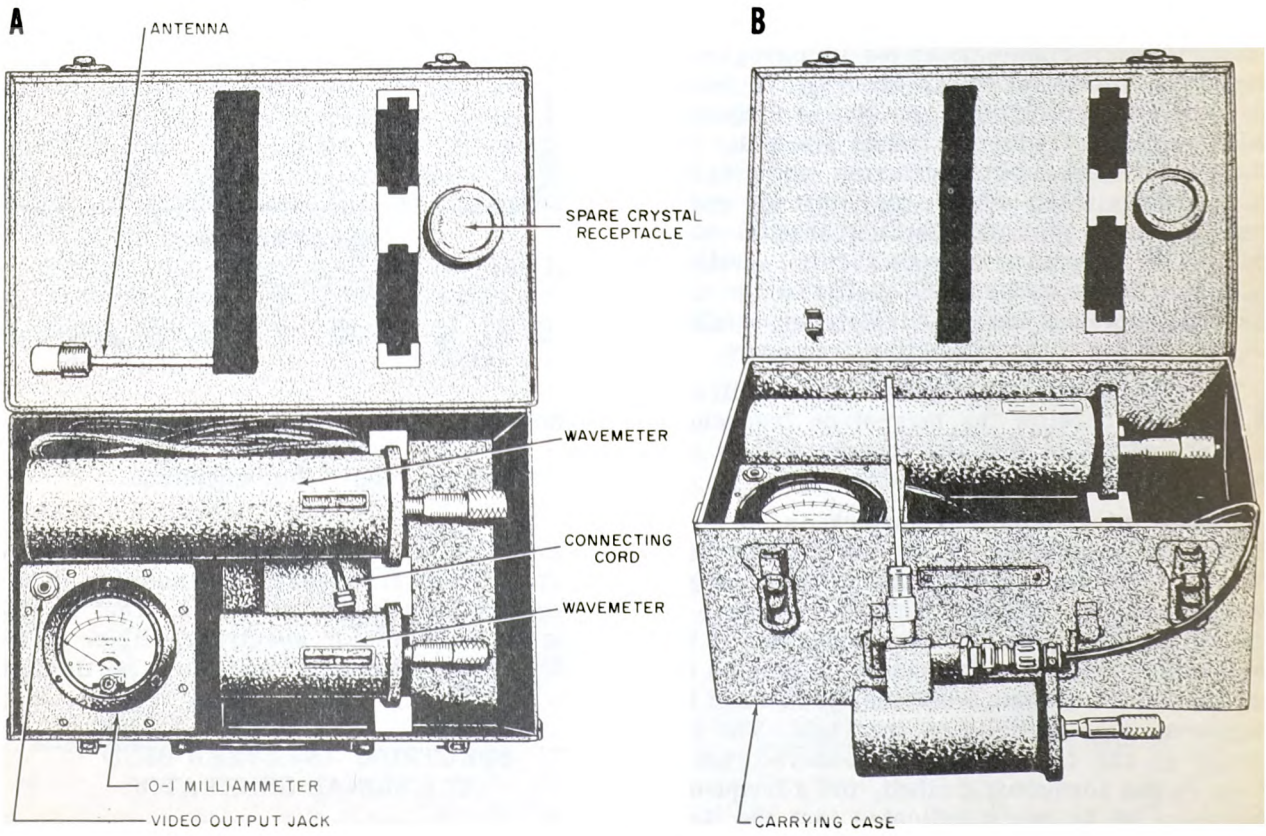

Figure 18-18.—Wavemeter test set.

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CIRCUIT OPERATION

Each of the wavemeters consists of a quarter-wave co-axial transmission line (fig. 18-19), which is capacitively tuned by a micrometer adjustment. A circuit diagram of the wavemeter is shown in figure 18-20.

From the study of transmission line theory, it should be recalled that for a given length of line having a given ratio of inner and outer conductor diameters, there is a frequency at which the section of the line will behave like a resonant circuit of low impedance. Also, by changing the capacitance of the line, as will be true when the micrometer (fig. 18-19) is moved in or out of the line, the equivalent length and resonant frequency of the section of line may be varied.

When an r-f signal is passed into the wavemeter from the antenna (fig. 18-20), a portion

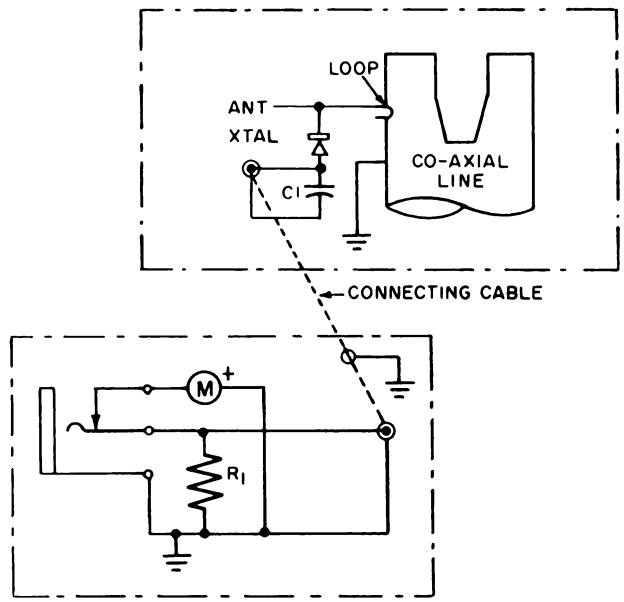

Figure 18-20.—Wavemeter circuit drawing.

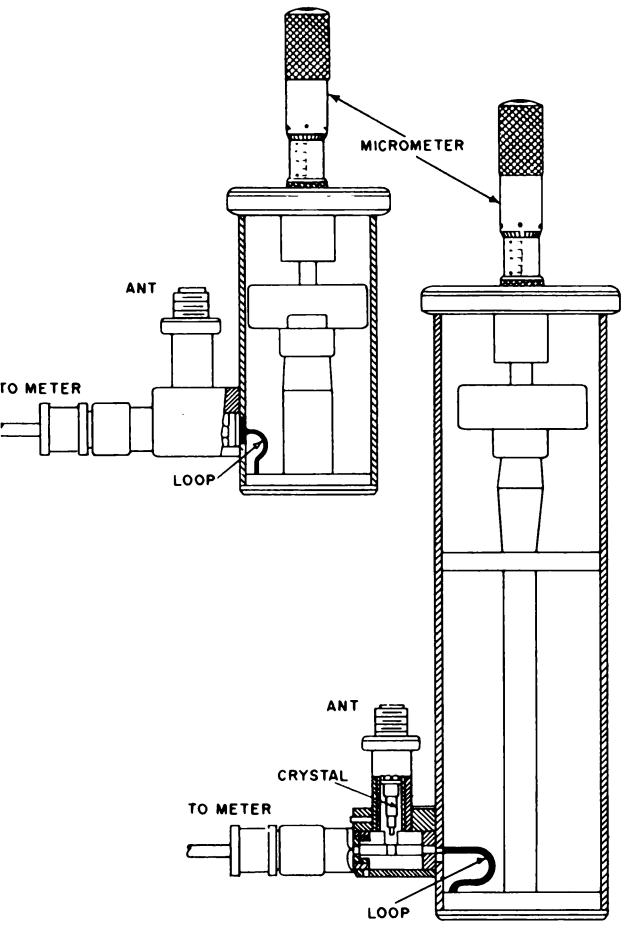

Figure 18-19.—Wavemeter longitudinal cross-section view.

of the energy is rectified by the crystal that passes through the cable to the milliammeter. The remainder of the r-f signal is fed into the co-axial line by a loop. When the wavemeter is tuned to resonance (by the tuning micrometer), the meter will show a dip because most of the energy is now being absorbed by the transmission line. The Q of the wavemeter is usually high enough to allow a sharp dip in the meter reading allowing a calibration accuracy within 1 mc.

The larger wavemeter of figure 18-19 covers the lowest frequency band from 80 to 360 mc. The tuning condenser consists of a fixed round plate and a variable flywheel plate. The variable plate is driven by the micrometer thimble.

The higher portion of the frequency band from 330 to 1220 mc is covered by the smallest wavemeter of figure 18-19. The internal construction of the two meters is the same.

READING THE MICROMETER DIAL

The graduations on the micrometer dial (fig. 18-21) are in two parts. The rotating thimble carries numbers from 0 to 24. A rotation to 25 will be the same as 0. The stationary barrel is engraved with a straight line along its length. This straight line is divided into 40 equal parts, each of which is equivalent to one rotation of the

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thimble. Each fourth dividing line is marked with a number. This number represents 100 scale divisions, and the lines between each number are equal to 25 scale divisions. To obtain the final micrometer reading multiply each subdivision on the shaft by 25 and add the number indicated on the product to the number obtained from the shaft.

A reading of 252 is indicated as the micrometer setting in figure 18-21. This is true since the number on the shaft is 2, representing 200; the number of divisions showing is two, representing 50; and the indication on the thimble is 2. The numbers are added to obtain the final reading of 252.

USE OF THE CALIBRATING TABLE

The sample calibration table (fig. 18-22) contains dial settings obtained from the micrometer. For each dial setting thus obtained, the table shows a frequency that corresponds to that setting. In this manner, the frequency of the source under test is determined.

The sample table shown is a specimen and does not indicate the actual calibration of a particular wavemeter test set. For reasons of simplicity, the sample table does not show decimal dial settings.

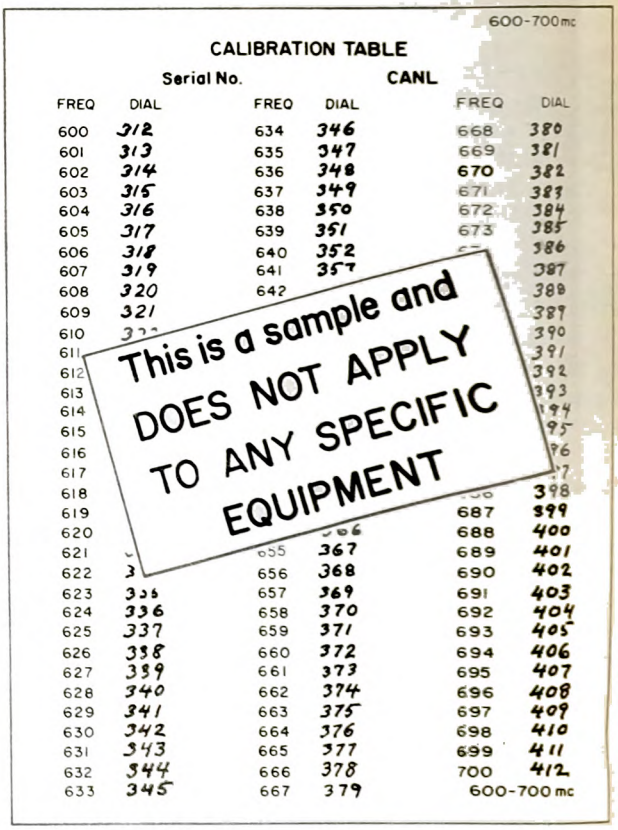

Figure 18-22.—Sample calibration table.

ELECTRONIC SWITCH

An electronic switch is a device that utilizes the properties of gas-filled or high vacuum tubes for closing, opening, or changing the operation of an electronic circuit. The electronic switch is faster and more sensitive than a mechanical switch. For example, an electronic switch may be used to gate an amplifier many times per second. This, in turn, causes the amplifier to function during a period of time and prevents its functioning at other times. The on and off periods of operation may have the same duration or they may be different, depending on the operating requirements.

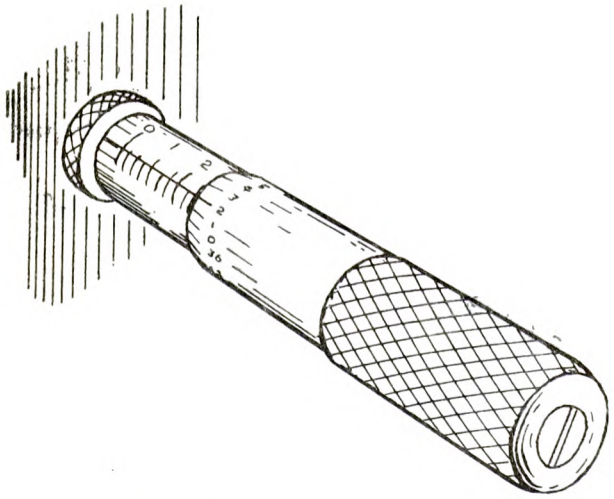

Figure 18-21.—Micrometer.

CIRCUIT OPERATION

A multivibrator (V2 and V3 of figure 18-23, A) may be used to cut the amplifier tube, V1, on and off at the multivibrator frequency. The waveforms of figure 18-23, B, indicate the manner in which the amplifier is gated.

The square-wave multivibrator output is developed across the cathode resistor of V3. It is indicated in figure 18-23, B, as the gate voltage, ek. When V3 conducts, its cathode becomes

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more positive, and the increase in potential is applied to the cathode of V1. Adding a positive potential to the cathode of V1 has the same effect as making the grid more negative. The rise in potential at the V3 cathode is sufficient to cut off V1. When V3 is not conducting, ek is zero, and V1 functions as an amplifier.

If the input voltage, e_{in}, to the amplifier tube, V1 consists of a series of regularly spaced positive pulses, (fig. 18-23, B), the gate voltage will block two of the pulses and allow two to pass. If the gate voltage existed for a longer period of time, three or more pulses would be passed and an equal number suppressed.

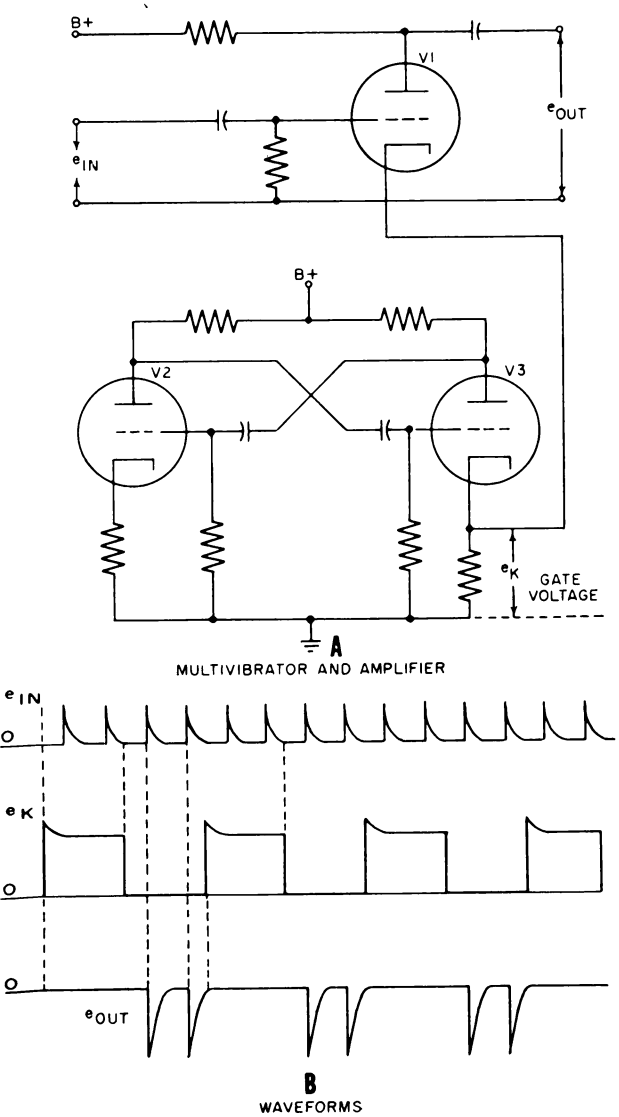

Figure 18-23.—Multivibrator used as an electronic switch.

APPLICATION

The switching action of a multivibrator may also be used to show, for purposes of comparison, two or more signals on a cathode-ray oscilloscope simultaneously. The frequency and phase relationship of the two signals may be determined from the simultaneous oscilloscopic display. Although the signals are not actually present at the same time, they appear to be so because of the persistence of vision of the human eye.

Consider the electronic switch circuit shown in figure 18-24. V1 and V2, with their associated circuits, make up a conventional multivibrator. V3 and V4 are separate amplifier stages having separate inputs but a common plate connection through C5 to J1. The output at J1 is connected to the vertical plates of the oscilloscope.

Assume that the input signal, B, is some multiple of the input signal, A, and, further, that the input signal, A, is used to synchronize the multivibrator. When V1 conducts, V3 is cut off and the amplified signal from V4 appears on the oscilloscope. When V2 conducts, V4 is cut off and the amplified signal from V3 appears on the oscilloscope. It is assumed that the time of one positive pulse from the multivibrator is equal to that of one or more periods of input signal A.

By connecting the output of the electronic switch to the vertical amplifier of the oscilloscope, the two signals are made to appear on the same horizontal axis of the oscilloscope screen. The output of V3 modulates one alternation of the square wave, and the output of V4 modulates the other. Adjusting the BALANCE control varies the output swing of the amplifiers and likewise the distance between the positive and negative alternations of the output. The gain A and gain B controls adjust the input amplitudes of the A and B signals, respectively. Through proper adjustment of the synchronizing controls on the oscilloscope, the two signals are made to start at the same point on the horizontal axis.

GRID DIP METER

The grid dip meter (fig. 18-25) is a vacuum tube oscillator to which a low-range milliammeter is added to read the oscillator grid current. The grid dip meter can be helpful to the Electronics Technician in determining the resonant frequency of deenergized tank circuits, determining the resonant frequency of antennas, or in checking field strength. Among its many other uses, the grid dip meter can be used as

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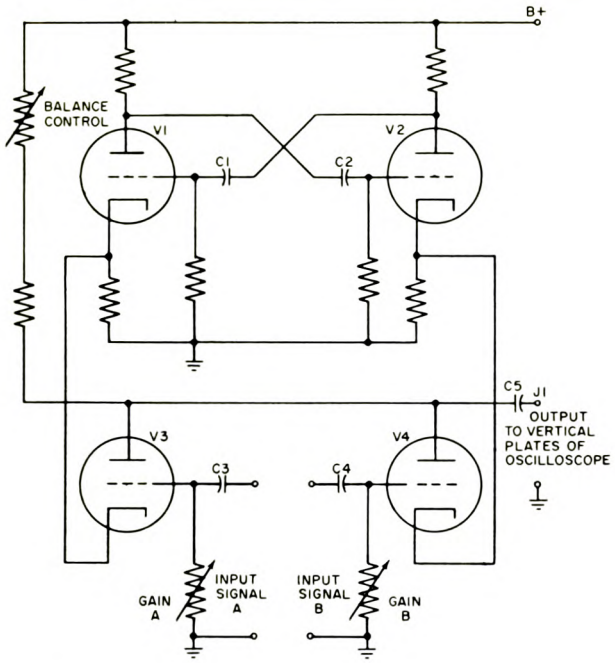

Figure 18-24.—Electronic switch for use with an oscilloscope.

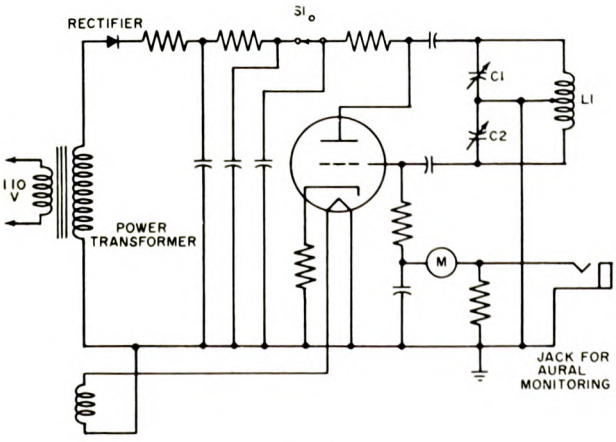

Figure 18-25.—Grid dip meter circuit diagram.

an absorption-type frequency meter when its oscillator is not energized.

CIRCUIT OPERATION

With the grid dip meter functioning, a small amount of energy is coupled from the tuned circuit, comprising L1, C1, and C2, to the circuit under test. Except for the field of L1, the circuit being tested is deenergized. The capacitors are rotated to bring the oscillator tank to the resonant frequency of the circuit under test. At the resonant point of the circuit, the grid current of the grid dip meter decreases, which will be indicated on the meter. The energy absorbed from L1 by the circuit being tested decreases the a-c component of the plate voltage, causing a decrease in feedback energy in the oscillator circuit and a drop in the grid current. A calibrated dial on the grid dip meter (not shown) indicates the frequency of resonance.

APPLICATIONS

In order to determine the resonant frequency of an antenna system, coil L1 of the meter is brought in close proximity to the antenna when the latter is deenergized. The proper point along the antenna is the point corresponding to a high current when the antenna is energized. If the antenna being tested has its center open, a jumper is used as a temporary short so that the test may be made. The capacitors are tuned until the dip of the meter indicates resonance. Harmonics of the antenna frequency may be indicated in the same manner.

Standing waves on an open transmission line may be checked by removing the plate power of the grid dip meter and operating the tube as a diode detector. In such cases, the grid to cathode rectification passes a current through the meter, which is proportional to the r-f component existing at that point. The operation is similar to that of a neon bulb indicator.

Parasitic oscillations may be detected through the use of the grid dip meter in the following manner. By closing S1, plate power is applied to the circuit. If, instead of observing the meter, a pair of headphones is inserted in the phone jack, an audible tone will be heard when the circuit is tuned by C1 and C2 through the fundamental or harmonic frequency of some r-f source. The frequency of the parasitic oscillation can be determined from the calibrated dial.

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QUIZ

. The spectrum analyzer presents on an oscilloscope display a pattern in which the relative amplitudes of the various frequencies of the spectrum are presented along what axis?

. When a carrier frequency is pulse modulated, what are the collective carrier frequency and its associated sidebands called?

. When a modulating frequency is heterodyned with a fundamental frequency, what is the relative amplitude of the sum and difference frequencies compared with that of the modulating frequency?

. To what frequency is the difference between any two adjacent frequencies equal in a pulsed radar spectrum?

. What is the effect of amplitude modulation (fig. 18-2, C) on a pulsed oscillator output spectrum?

. What is the effect of frequency modulation (fig. 18-4, C) on a pulsed oscillator output spectrum?

. What is the waveform of the r-f oscillator modulating voltage in the spectrum analyzer (fig. 18-5)?

8. What general class of r-f oscillator is used in the spectrum analyzer of figure 18-5?

9. What is the frequency range of the sweep oscillator (fig. 18-5)?

10. What is the purpose of frequency modulating the r-f oscillator V8 in figure 18-6?

11. The detector stage, V3 (fig. 18-6) operates as an infinite impedance detector for what purpose?

12. The grids of V10 (fig. 18-6) are connected in what type of circuit?

13. When selector switch S1 (fig. 18-6) is in the MIXER position, the V11B grid is grounded for what reason?

14. All power for absorption type wavemeters is obtained from what source?

15. Adjusting the balance control (fig. 18-24) has what effect on the amplifier output?

16. When a grid dip meter is used as an absorption type frequency meter what is the condition of its oscillator?

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APPENDIX I

ANSWERS TO QUIZZES

Chapter 1

INTRODUCTION TO ET 2

1. Yes.
2. (1) Communications ET, (2) Radar ET, and (3) Auxiliary and Miscellaneous ET.
3. Technical.
4. Preventive.
5. Read basic material.
6. EIB.
7. Record of Practical Factors.
8. Record of Practical Factors.
9. Service, Fleet, and Manufacturers.
10. On-the-job.
11. On-the-job.

Chapter 2

ELECTRONIC AIDS TO NAVIGATION: RDF (PART I)

1. (a) A sensitive receiver to which a directional antenna is connected. (b) Visual indication of the bearing of a received signal.
2. The antenna, receiver group, and azimuth indicator.
3. Vertically polarized Adcock array, which is motor driven at a constant speed.
4. A propeller-shaped pattern.
5. Balan.
6. (a) Variable frequency oscillator. (b) A tuning capacitor and a reactance tube shunted across the vfo.
7. 4th to the 13th harmonic.
8. Low band-pass filter.
9. Frequency quintupler and frequency doubler.
10. Five frequencies.
11. Two frequencies.
12. Phase deviation.
13. Motor control amplifier and reactance tube.
14. Memory cylinder.
15. Rotating capacitances C3 and C4.
16. 900 rpm.
17. R-f head.
18. An inductance, L.
19. Series-fed Hartley.
20. 15.3 mc.
21. The unbalance of bridge circuits.
22. Monitor circuit.
23. The indicator and monitor.
24. C10 through C19 in conjunction with C20 and C21.
25. 5.172917 or 5.181250 mc.
26. To lower the Q of the circuits.
27. CP2 and CP3.
28. An output voltage proportional to phase.
29. Clutch control amplifier and discriminator circuits.
30. Determine the direction of rotation of the motor.
31. The high-, medium-, and low-speed clutches, respectively.
32. Motor control amplifier V13.
33. Discriminator.
34. +5 volts.
35. Zero volts.
36. Reactance tube.
37. A lower frequency.
38. From the high to the low frequency end.
39. Tuning motor B2 (fig. 2-10), the drive mechanism, and the selector motor, B1 (fig. 2-15).

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Chapter 3

ELECTRONIC AIDS TO NAVIGATION: RDF (PART II)

1. Amplify and detect the i-f signal.
2. AGC amplifier, V31.
3. +21 volts.
4. AGC rectifier, V32A.
5. Delayed AGC.
6. R-f gain control, R250.
7. Compensates for losses in transformer coupling at high frequencies.
8. Audio level control, R246.
9. Squelch tube, V34A.
10. To remove the control of the squelch circuit.
11. 150 to 4500 cps.
12. Shorting of the T1 secondary to mute the audio output.
13. The indicator chassis.
14. The arm of R2.
15. A transitron circuit.
16. Balance controls.
17. 90 degrees.
18. -150-volt and +275-volt supplies.
19. The vertical and horizontal 15-cycle alternator inputs which are 90° out-of-phase.
20. By rectifying an r-f voltage generated by an oscillator, V16.

Chapter 4

OPERATING PRINCIPLES OF A REPRESENTATIVE RADIO RECEIVER

1. Ground waves.
2. Medium-frequency band.
3. Short antennas.
4. Dielectric constant.
5. (a) 14 kc to 32 mc. (b) Five.
6. Two.
7. (a) 60 kc. (b) 200 kc.
8. The filter and the first i-f amplifier.
9. A frequency converter.
10. Calibrating the main tuning dial.
11. The band selector.
12. Tuning capacitor C5.
13. R-f amplifier section.
14. Long-wire or whip type.
15. S1A.
16. A filter that prevents r-f variations from entering the biasing circuit.
17. (a) Electron-coupled Hartley oscillator. (b) C26.
18. (a) 60 kc. (b) 200 kc.
19. Electron-coupled Pierce circuit.
20. 59 to 61 kc.
21. Contact 1 for FSK, contact 2 for A1 broad, and contact 3 for A1 sharp.
22. (a) Antenna preamplifier and r-f amplifier. (b) The first and second i-f amplifier of the second i-f assembly.
23. Temperature compensating resistor.
24. A1 BROAD, A1 SHARP, or FSK position.
25. C42.
26. S2E of the reception control.
27. An instantaneous decrease in potential at the noise limiter plate.
28. Add decibel switch, S4.
29. A 50-kc Pierce oscillator comprising V22 and a 10-kc multivibrator, V23.
30. (a) Crystal Y2 and capacitor C87. (b) C90.
31. The 50-kc signal from the cathode of V22.
32. (a) The r-f assembly (fig. 4-4). (b) The second i-f amplifier of the second i-f assembly.
33. The plate of the voltage regulator tube, V26, to ground.
34. The crystal calibrator.
35. A crystal-controlled, cathode-coupled multivibrator.
36. The band selector, section S1D.
37. All bands.
38. The sharp filter, Z2.
39. AGC diode, V14.
40. (a) R-f amplifiers. (b) First and second i-f amplifiers.
41. F1 and A3 reception.
42. The slave receiver.
43. A3 sharp or A3 broad.

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Chapter 5

OPERATING PRINCIPLES OF A REPRESENTATIVE RADIO TRANSMITTER, PART I

1. 0.3 to 26 mc.
2. (1) Voice, (2) facsimile, (3) c-w, and (4) frequency shift telegraph communication.
3. Manual.
4. 500 watt.
5. Radio frequency oscillator.
6. 2 to 26 mc.
7. 14 subunits.
8. Units 2, 5, 6, and 8.
9. Units 4, 5, 6, 7, and 10.
10. (a) Unit 5 and (b) unit 6.
11. (a) 10.4 to 10.5 mc and (b) unit 8.
12. A 5-mc and a 15-mc signal respectively.
13. Unit 14.
14. (a) A crystal oscillator and four cathode followers and (b) the cathode followers.
15. A frequency quintupler and a frequency doubler.
16. 90 to 100 kc.
17. Frequency shift keying and facsimile.
18. The oscillator frequency increases.
19. Unit 5.
20. Unit 1.
21. (a) Five and (b) 1 mc from unit 2, 100 kc from unit 1, 100 kc $\pm\Delta$ from unit 12, 90 to 100 kc from unit 3, and an 8-mc signal from unit 4.
22. To synchronize the step oscillator, V3.
23. One tenth of the isolating amplifier output frequency or 1/10 the crystal oscillator frequency.
24. C3 and R3.
25. Crystal CR1.
26. Control grid of amplifiers V2 and tripler V3.
27. (a) Unit 9 and (b) unit 8.
28. The 15-mc output of unit 7.
29. 27 to 28 mc input from unit 8 and a 5-mc signal from unit 7.
30. Either unit 11A, 11B, or 11C.
31. Position 3.
32. Bandswitch located in unit 14.
33. (a) Unit 11A, (b) unit 11B, and (c) unit 11C.
34. Mixer V1.
35. To provide the high power output required from the cathode follower.
36. (a) 13, 14, 16, 17, 18, 19, 20, and 22 mc and (b) 22 to 23, 27 to 28, 32 to 33 or 37 to 38 mc.
37. Nine.
38. (a) To the megacycle value and (b) to the second figure of the megacycle value.
39. The mixer, V1, and the first amplifier, V2.

Chapter 6

OPERATING PRINCIPLES OF A REPRESENTATIVE RADIO TRANSMITTER, PART II

1. 0.3 to 26 mc.
2. The buffer, the intermediate power amplifier, and the power amplifier.
3. (a) Excitation control, R1.
   (b) 5 volts rms.
4. Bandswitch sections S2A, S2B, and S2C.
5. Capacitor C3.
6. Capacitor C13.
7. Inductor L15 and capacitor C14.
8. 300 volt.
9. 500 watt.
10. (a) Bandswitch control.
    (b) Bands 1 and 2.
11. Keying relay K6.
12. Standing-wave ratio monitor and the impedance transformer circuits.
13. 4:1 for any frequency in the transmitter range.
14. Loading switch.
15. Antenna transfer switch.
16. (a) The transformer switch, S1.
    (b) Step up the signal to the required level.
17. Impedance transformer.

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Chapter 7
OPERATING PRINCIPLES OF A REPRESENTATIVE RADIO TRANSMITTER, PART III

1. (a) Units 11A, 11B, and 11C and (b) RFO.
2. Handset jack J1 or microphone jack J2.
3. Less AGC voltage is required to produce a highly sensitive AGC circuit.
4. V3.
5. Multivibrator V7, crystal rectifier CR1, and limiter V6B.
6. Handset receptacle J1, sidetone jack J4, and a remote radiophone unit via J3.
7. So that the cathode output will not be of a greater amplitude than that of the plate.
8. Push-pull power amplifiers V9 and V10.
9. -40 volts.
10. Keying relay K6.
11. Assures that V13B conducts and V13A is cut off during space.
12. Push-pull cathode follower.
13. 10F and 1F of S4B front.
14. 250 watts.
15. Transformer T1.
16. +3000 volts.
17. MVPS, RFO, LLRM, RFA, and the antenna coupler.
18. R1 and C1.
19. The motor and clutch of the time delay relay, K2.
20. Double-choke input filter.
21. 100 watt.
22. -24 V after time delay line.
23. AN/SRT-15 and 16.
24. Booster emergency switch S1.
25. 500 W disable line (fig. 7-8).
26. C-w, frequency-shift, and facsimile operation.
27. +2400 volt.

Chapter 8
PRINCIPLES OF SINGLE-SIDEBAND COMMUNICATIONS

1. The power in the sidebands is equal to one-half that in the carrier.
2. 100 watts.
3. One-half.
4. The permissible frequency variation is $\frac{1 \text{ th}}{100}$ of that for an a-m system.
5. 1 kc.
6. To produce lower and upper sideband frequencies of 249 and 251 kc and to suppress the 250-kc oscillator frequency.
7. To pass the upper sideband (251 kc) and suppress the lower sideband (249 kc).
8. To pass only 251 kc.
9. After the mechanical filter at the input of the second balanced modulator.
10. Any conventional a-m receiver that is capable of being tuned to this signal.
11. The crystal frequency is equal to the nominal carrier frequency plus 1400 kc.
12. Speech clarifier.
13. Capacitor C discharges through diode 2.
14. The filter method and the phasing method.
15. They are in phase.
16. The degree of balance that is maintained in the plate circuits of V1 and V2.
17. 180 degrees out of phase.
18. The lower sideband from balanced modulator 1 and the lower sideband from balanced modulator 2.
19. Because the unwanted sideband is phased out in the balanced modulator.
20. Crystal filters have a Q of 10,000 to 200,000 or higher; mechanical filters have a Q of 2,000 to 10,000 or more.
21. The frequency is constant even during wide temperature changes.
22. A bridge circuit.
23. From 350 to 3000 cycles.
24. Amplitude modulated, with suppressed carrier and the upper and lower sidebands utilized as twin channels (conventional double sideband also available).
25. By wobbling the carrier frequency plus and minus 250 cycles from the mean value.
26. The r-f amplifier.
27. Because its oscillator frequency (1700 kc) is below the frequency (1750 kc) of the incoming signal.
28. To separate the sidebands from the carrier.
29. Signal comparator (2).
30. Heterodyne.
31. To hold the capacitor drive motor locked in the absence of an incoming signal.
32. Signal comparator (1)

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Chapter 9
BASIC PRINCIPLES OF COMMUNICATIONS MULTIPLEXING

1. Multiplex transmission.
2. Slightly wider than n times that necessary for one channel.
3. The sum of the bandwidths of the individual channels, plus the sum of the necessary guard-band frequencies between channels.
4. Pulse amplitude, pulse width, or pulse position.
5. 3,875 cps.
6. 107 μ s.
7. Channel selectors.
8. 4.75 μ s.
9. To act as a synchronizing voltage between the receiving and transmitting multiplex terminals.
10. Sawtooth oscillator V1 and the discharge stage, V4.
11. Marker pulse generator.
12. 2.2μs.
13. C30.
14. 2.2μs.
15. The degree of synchronization that can be maintained between the multiplex and demultiplex equipments.
16. Sawtooth trigger pulse in the multiplex waveform generator.
17. 40 volts.
18. Channel 6.
19. The pulse position.
20. 5.2μs.

Chapter 10
RADIO TELETYPE SYSTEM, PART I

1. Coded.
2. Send-receive.
3. A frequency above its assigned or rest frequency.
4. 1 to 6.7 mc.
5. Switches the teletypewriter to any one of three d-c loops.
6. Diversity.
7. Two selenium rectifiers, CR1 and CR2.
8. Increase.
9. Terminals 9 through 12 of T1.
0. Interconnection and transfer of teletype equipment with various radio adapters.
1. The teletype equipment to the looping jacks.
2. (a) Shifting a constant amplitude carrier between two fixed frequencies (b) Marking.
13. Frequency-shift oscillator.
14. Balanced reactance modulator.
15. Balanced mixer circuit.
16. 6 watts into a 75-ohm, noninductive resistive load.
17. Pulse shaper.
18. 8 volts.
19. Frequency multiplication in the transmitter.
20. The reactance modulator (fig. 8-11).
21. 0 to 1000 cycles.
22. 0 and 1.
23. (a) Three bands. (b) 1 to 1.8 mc, 1.8 to 3.5 mc, and 3.5 to 6.7 mc.
24. Cathode balancing adjust R91.
25. S2B.

Chapter 11
RADIO TELETYPE SYSTEM, PART II

1. 60 microwatts to 60 milliwatts.
2. 600 ohms.
3. Improve the balance between mark and space signals.
4. Frequency-cps switch, S4.
5. Keyer circuit.
6. To open and close the d-c loop of the teletype printer.
7. Automatically comparing the two diversity signals.
8. 700 cps for marking signals, and 500 cps for spacing signals.
9. Corresponding d-c voltage.
10. Receive bias control, R28.
11. Marking bias.
12. (a) 60 ma. and (b) Zero.
13. Causes the terminal to switch from standby to receive.

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Chapter 12

COMMON MAINTENANCE PROBLEMS: RADIO AND RADIAC

1. POMSEE.
2. Performance, Operational, and Maintenance Standards for Electronic Equipments.
3. Binder for Electronic Equipment Performance Standards Sheets.
4. No.
5. Performance Standards Sheet.
6. Against themselves.
7. The sliding short on the main tuning coil of the r-f tuner.
8. A serious mismatch will exist.
9. A serviceable replacement drawer may be substituted.
10. -30 volts.
11. By connecting one of the 100-kc Xtal outputs to the X axis of the oscilloscope, and the other to the Y axis.
12. Inductance L1.
13. It throws the unit out of alignment.
14. 10-kc jack.
15. Unit 6.
16. To check the r-f input and output, respectively.
17. Dummy load DA-91/U.
18. 5 volts.
19. The peak at the lowest tune IPA dial reading.
20. 50.
21. A SWR lower than the chosen position of the SWR calibrate control.
22. Audio, keying, and power supply circuits.
23. Stage gain measurements.
24. Testing subminiature tubes mounted on the subassemblies.
25. So as to connect the elements of the tubes in the subassemblies being tested to the corresponding circuits of the tube tester.
26. Testing of type 5644 voltage regulator tubes.
27. -40 to +85 degrees centigrade.
28. Cross-mat parallel wiring.

Chapter 13

ELECTRONIC AIDS TO NAVIGATION: LORAN

1. Measurements of very small periods of time.
2. By using four different transmitting frequencies, and by using different pulse recurrence rates.
3. Channel 1, 1950 kc; channel 2, 1850 kc; channel 3, 1900 kc; and channel 4, 1750 kc.
4. Channel and PRR switches (S3 and S6, respectively).
5. Local-distant switch S2.
6. Low-frequency components.
7. The received signals.
8. AFC switch, S5.
9. V8 through V14.
10. An 80-kc, sine-wave oscillator, V17.
11. One-half the frequency of the input.
12. Repetition rate and variable delay circuits.
13. When the predetermined count selected by the PRR switch, S6, is reached.
14. Twice that of the received signals.
15. Pedestal circuits.
16. The B pedestal.
17. On the A and B traces, respectively.
18. Allow the circuit to be resonated with antennas having different characteristics.
19. Either 20 or 40 db.
20. Pierce oscillator circuit.
21. A clipper and amplifier stage.
22. Reactance tube, V16.
23. Whether the received signals are faster or slower than the receiving set repetition rate.
24. Reactance tube, V16.
25. The counter chain.
26. From V22 (the shortest negative pulse).
27. The lower B pedestal.
28. A cathode-coupled multivibrator.
29. Fixed delay circuits.
30. The reset thyratron plate.
31. The master and slave pulses during the A and B periods, respectively.

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Chapter 14

ELECTRONIC AIDS TO NAVIGATION: TACAN (PART I)

1. An air navigation system called TACAN.
2. Receiver-transmitter group, antenna group, and power supply test set group.
3. 100 aircraft.
4. 63 mc.
5. Beacon identification, distance information, bearing information, and random pulses used to make up a constant duty cycle.
6. Exactly 50 μs.
7. The antenna.
8. To establish magnetic north as a reference point in time.
9. Azimuth information.
10. To improve bearing accuracy.
11. Distance-measuring information.
12. The number of pulses passing through the multivibrator (during an interrogation period) to the receiver output.
13. The upper antenna pulser coil.
14. 24 cycles.
15. 3600 pulse-pairs with 12 μs spacing between pulses of a pair.
16. By the distance (in time) between the two delayed pulses from the second delay line.
17. They develop a larger output voltage with good output regulation, despite variations in load impedance.
18. 40 to 51 mc.
19. 3.5 μs.
20. So that the 3.5-μs modulating pulse will arrive at approximately the center time of the klystron beam pulse and prevent undesirable distortion of the output waveshape.
21. 6 kw.
22. A frequency sensitive device, a preamplifier, with one saturable reactor, and a power amplifier with three saturable reactors.
23. To increase the effective capacitance in the correction circuit.
24. To the zero signal position.
25. The antistickoff voltage.
26. To make the fixed phase in quadrature with the control phase to produce starting torque when the control phase is energized.
27. The -375- and -400-volt supplies are in operation.
28. Low voltage power supply.

Chapter 15

ELECTRONIC AIDS TO NAVIGATION: TACAN (PART II)

1. Duplexer.
2. 1025 and 1087 mc on the low band and 1088 and 1150 mc for high band reception.
3. Tickler coil.
4. 40 to 50.5 mc.
5. 962 and 1024 mc or between 1151 and 1213 mc.
6. To avoid the possibility of beat frequencies between the receiver local oscillator and transmitter.
7. 120.35 to 151.6 mc.
8. To limit the signal amplitude applied at the grid.
9. 2 mc.
10. V10, V12, and V13.
11. V9 and V10.
12. 2700 ± 90 pps.
13. 12-μs delay line, W1.
14. 12 μs.
15. A single negative pulse.
16. 40 us.
17. Squitter control voltage.
18. D-c clamp diode, V23.
19. The radio receiver and antenna reference burst generator.
20. One-fourth.
21. 135 cps pulse gate multivibrator, V3A.
22. The period of the reference bursts.
23. Phase-locking.
24. One output pulse.
25. 3600 pulse-pairs spaced 12 μs apart.
26. The distance in time between the leading edges of the two pulses from the delay line, DL2.
27. V12 and V13.
28. Klystron, V15.
29. Oscillator chassis, the pulse shaper chassis, and three coaxial amplifiers.
30. A directional coupler.
31. The klystron beam pulse must be present at the cathode.

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Chapter 16

ELECTRONIC AIDS TO NAVIGATION: TACAN (PART III)

1. The spacing of the soft iron slugs about the outer edge of the pulser plate.
2. A synchro system.
3. Data that corresponds to the angular position of its rotor.
4. It is equipped with an inertial damper.
5. 90° from the rotor of the transmitter synchro.
6. A voltage indication of the error in the angular positions of the input and output shafts of a servomechanism.
7. The direction in which the rotor of the control transformer is out of correspondence with the rotor of the transmitter.
8. The receiver rotor with respect to the transmitter rotor.
9. The 1- and 36-speed differential generators.
10. The bearing servo motor.
11. A reference voltage.
12. The rate of change of the shaft position.
13. 2.5 v.
14. False nulls.
15. 2.5°.
16. To control the energized and deenergized condition of switching relay K2.
17. Push-pull.
18. A phase sensitive detector.
19. Movements of the ship.
20. Changes the tachometer voltage from rate feedback to acceleration feedback.
21. Power amplifier 3-2 load winding.
22. 100-watt, two-phase motors, having a continuously excited phase and a variable phase.
23. 900 rpm.
24. T1, Z1, and Z2.
25. Series.
26. The information is the same.
27. AN/SRN-6.
28. 7.5 kw.
29. A pair of tunable resonant cavities.
30. A 675-cps signal.
31. The pulse-sweep generator.

Chapter 17

THE USE OF SPECIAL TEST EQUIPMENT, PART I

1. 20 to 400 mc.
2. Gain of the i-f section can be adjusted.
3. VTVM stage.
4. Peak value.
5. Slope detection.
6. Bandwidth.
7. Volts-per-meter.
8. A 50-ohm resistive load.
9. Above the range of the RI-FI meter.
10. 2, 4, 8 and 6.
11. To control the gain of the first i-f amplifier, V1, and the entire i-f section.
12. By the different position settings of the function switch, S1.
13. The impulse calibrator, Z3.
14. Quasi-peak.
15. Difference in the value of load resistance.
16. One-shot multivibrator.
17. Balanced bridge circuit.
18. CAL position.
19. Input attenuator.
20. Marker interval control.
21. Trigger generator.
22. When a very high impedance must be presented to the circuit under test.
23. 0.35 μs.
24. Voltage calibrator.
25. Cd (stray capacitance).
26. 45 to 5500 PPS.
27. Blanking marker.

Chapter 18

THE USE OF SPECIAL TEST EQUIPMENT, PART II

1. The vertical, or Y axis.
2. A spectrum.
3. The amplitudes of the sum and difference frequencies are each half the amplitude of the modulating frequency.
4. The fundamental modulation frequency.
5. It increases the number of sidebands in the spectrum.
6. It increases the amplitude of the side lobes.
7. Sawtooth.
8. A reflex velocity-modulated klystron.
9. 10-30 cps.
10. To display on the scope screen each frequency component of the signal under test.
11. To minimize the loading effect on the tuned tank L3-C13.
12. A balanced bridge.
13. To prevent i-f signals and the frequency meter pip from being intensified.
14. From the circuit being tested.
15. It varies the output swing.
16. Not energized.

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APPENDIX II

QUALIFICATIONS FOR ADVANCEMENT IN RATING

ELECTRONICS RATINGS
ELECTRONICS TECHNICIANS (ET)

Quals Current Through Change 15

General Service Rating

Scope

Electronics technicians maintain, repair, calibrate, tune, and adjust all electronic material—except airborne and weapon control equipment—used for communication—except interior communication systems and teletypewriter machines—detection and tracking, recognition and identification, aids to navigation, and electronic countermeasures.

Emergency Service Ratings

ELECTRONICS TECHNICIANS N (Communications) . . . . . ETN
Maintain, repair, calibrate, tune, and adjust communication equipment, radio aids to navigation, and radio countermeasures equipment including: Radio equipment; facsimile equipment; teletype terminal equipment and similar types of terminal equipment; data transmission systems; radio direction finding equipment; loran receiving equipment and radio beacons.

ELECTRONICS TECHNICIANS R (Radar) . . . . . ETR
Maintain, repair, calibrate, tune, and adjust electronic sea, land, and air detection and tracking equipment; electronic recognition and identification equipment and radar countermeasures equipment including: Search radar equipment, radiac equipment, IFF systems, and racons.

ELECTRONICS TECHNICIANS S (Sonar) . . . . . ETS
Maintain, repair, calibrate, tune, and adjust electronic underwater detection and tracking equipment and sonar countermeasures equipment.

Navy Enlisted Classification Codes

For specific Navy enlisted classification codes included within this rating, see Manual of Navy Enlisted Classifications, NavPers 15105 (Revised), codes ET-1500 to ET-1599.

Throughout quals asterisk(*) denotes items which, when applied to emergency service ratings, are limited to the equipment, tools, and test equipment included within the scope of those ratings.

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Qualifications for Advancement in Rating	APPLICABLE RATES			
	ET	ETN	ETR	ETS
100 PRACTICAL FACTORS				
101 OPERATIONAL				
1. Manipulate external controls; read and interpret dials, meter indications, and cathode ray tube presentations on electronic equipment*	3	3	3	3
2. Utilize distribution patching system for radio transmitters, receivers, and antennas for all local and remote operating positions	3	3	--	--
102 MAINTENANCE AND/OR REPAIR				
1. Demonstrate under simulated conditions the rescue of a person in contact with an energized electrical circuit, resuscitation of a person unconscious from electrical shock, and treatment for electrical and acid burns.	3	3	3	3
2. Demonstrate, while servicing equipment, safety precautions such as tagging switches, removing fuses, grounding test equipment, and using shorting bars and rubber mats	3	3	3	3
3. Read schematic and wiring diagrams of electrical and electronic circuits; identify and interpret electrical, electronic, and mechanical symbols shown in electronic technical maintenance publications and installation blueprints	3	3	3	3
4. Locate and identify components, assemblies, subassemblies, and primary and casualty power circuits of electronic equipment by reference to technical maintenance publications, block diagrams, and installation blueprints*	3	3	3	3
5. Interpret RETMA color coding of: Fixed capacitors and resistors; power, AF, and IF transformer connections	3	3	3	3
6. Inspect, clean, and lubricate electronic equipment in accordance with technical maintenance publications*	3	3	3	3
7. Select, use, and maintain hand tools and small portable power tools necessary for maintenance and repair of electronic equipment*	3	3	3	3
8. Demonstrate use of the following test equipment:				
a. Nonelectronic multimeter	3	3	3	3
b. Electronic multimeter	3	3	3	3
c. Tube tester	3	3	3	3
d. Oscilloscope	3	3	3	3
e. Capacitance-inductance-resistance bridge	3	3	3	3
f. AF signal generator	3	3	3	3
g. RF signal generator	3	3	3	3
h. Frequency standards	3	3	3	3
i. Megohmmeter	3	3	3	3
j. Range mark generator	3	--	3	3
k. Echo boxes	3	--	3	--

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Qualifications for Advancement in Rating	APPLICABLE RATES			
	ET	ETN	ETR	ETS
102 MAINTENANCE AND/OR REPAIR—Continued				
9. Perform operational test and make external adjustments on electronic equipment*	3	3	3	3
10. Make tests for short circuits, grounds, and continuity of interconnecting cables between components of electronic equipment*	3	3	3	3
11. Make electrical connections and splices including soldered joints	3	3	3	3
12. Test and/or replace plugs, lamps, fuses, switches, electron tubes, jacks, cables, and wiring	3	3	3	3
13. Localize equipment casualties to components of a system of electronic equipment*	3	3	3	3
14. Inspect and clean commutators and slip ring assemblies; inspect and replace brushes	3	3	3	3
15. Take sonar beam patterns and perform range and bearing accuracy tests	2	--	--	3
16. Operate the following test equipment:				
a. Electronic switch	2	2	2	2
b. Synchroscope	2	--	2	2
c. Absorption wavemeter	2	2	2	--
d. Grid dip meter	2	2	2	--
e. Radio-interference field-intensity meter	2	2	2	--
f. Spectrum analyzer	2	2	2	--
17. Repair multimeters, oscilloscopes, test oscillators, and signal generators	2	2	2	2
18. Test electronic circuits for continuity, short circuits, and grounds; measure electrical quantities such as voltage, current, power, and frequency and compare with established values; use an oscilloscope to view circuit waveforms and compare with established optimum performance waveforms required in electronic equipment*	2	2	2	2
19. Perform tests, adjustments, and repairs necessary for proper operation of electromechanical servomechanisms and synchro control circuits including:				
a. Electrical zeroing of synchros	2	2	2	2
b. Testing servomotors and amplidynes	2	2	2	2
c. Gain, phase, and balancing adjustments	2	2	2	2
20. Perform sensitivity and selectivity measurements and aline circuits as necessary for optimum performance of electronic equipment*	2	2	2	2
21. Localize electronic equipment casualties to parts or subassemblies; repair by replacement of subassemblies or parts*	2	2	2	2
22. Effect authorized field changes to electronic equipment in accordance with instructions and diagrams*	2	2	2	2
23. View and compare with established standards, waveforms of the following circuits: Squaring and peaking, clamping circuits, high vacuum tube sweep generators (hard tube type), trapezoidal sweep generator, phantastrons, blocking oscillators, and counting circuits	1	--	1	--

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Qualifications for Advancement in Rating	APPLICABLE RATES			
	ET	ETN	ETR	ETS
102 MAINTENANCE AND/OR REPAIR—Continued				
24. Aline circuits by synchronizing multivibrators or blocking oscillators with sine wave, positive pulses, or submultiples of trigger frequency, or negative pulses	1	--	1	--
25. Evaluate test equipment for correct operation; make authorized repairs and calibrations	1	1	1	1
26. Repair countermeasures and radiac equipment*	1	1	1	1
27. Analyze and evaluate electrical and electronic tests; make adjustments, calibrations, and repairs necessary for optimum performance of electronic equipment*	1	1	1	1
28. Adjust for traffic requirements or environmental conditions antenna arrays such as: Driven arrays (collinear), parasitic arrays (Yagi), parabolic or corner or flat reflectors back of dipoles, phased arrays, waveguide, type antennas	C	C	C	--
29. Test and evaluate for proper and secure installation and optimum performance, newly installed or overhauled components, assemblies, or subassemblies of electronic equipment*	C	C	C	C
103 ADMINISTRATIVE AND/OR CLERICAL				
1. Locate in technical and maintenance publications information necessary for maintenance and repair of electronic equipment; enter corrections to publications when changes are made	3	3	3	3
2. Record test data and work accomplished in required work logs, equipment histories, and checkoff lists	3	3	3	3
3. Take, record, and report inventories of tools and portable test equipment available for maintenance and repair of electronic equipment	3	3	3	3
4. Gather information for and prepare and complete entries in electronic equipment failure reports	2	2	2	2
5. Obtain part and stock numbers from technical and supply publications for tools and replacement parts; procedures for requisitioning such material	2	2	2	2
6. Prepare job orders and work requests	2	2	2	2
7. Supervise and train personnel in maintenance and repair of:				
a. Radio equipment	1	1	--	--
b. Radar equipment	1	--	1	--
c. Sonar equipment	1	--	--	1
d. Countermeasures equipment	1	1	1	1
8. Prepare and complete periodic or recurring reports concerning performance and/or maintenance of electronic equipment	1	1	1	1

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Qualifications for Advancement in Rating	APPLICABLE RATES			
	ET	ETN	ETR	ETS
103 ADMINISTRATIVE AND/OR CLERICAL—Continued				
9. Inspect completed work logs and checklists; review electronic equipment failure reports, requisitions for tools and replacement parts, inventories of tools and portable test equipment, job orders, and work requests	1	1	1	1
10. Qualify individual operators to carry out authorized performance standards tests on electronic equipment to which assigned	C	C	C	C
11. Organize and administer maintenance and repair of electronic equipment	C	C	C	C
12. Evaluate completed electronic equipment failure reports, requisitions for tools and replacement parts, inventories of tools and portable test equipment, job orders, work requests, electronic equipment histories, and periodic or recurring reports concerning performance and/or maintenance of electronic equipment	C	C	C	C
200 EXAMINATION SUBJECTS				
201 OPERATIONAL				
1. Purpose of operator's controls and adjustments such as:				
a. Receiver gain, transmitter tuning, antenna tuning, loran sweep speed switch, and coarse and fine delay	3	3	--	--
b. Intensity, focus, receiver tuning, antenna rotation, range, and IFF interrogation switch	3	--	3	--
c. Receiver gain, intensity, training, keying selector, and trace sensitivity	3	--	--	3
2. Purpose of cathode ray tube presentations on electronic equipment*	3	3	3	3
3. Purpose of:				
a. Stages of a typical radio transmitter—master oscillator, modulator, power amplifier, and antenna	3	3	--	--
b. Stages of a typical superheterodyne receiver—antenna, RF amplifier, oscillator, mixer, IF amplifier, detector, AF amplifier, and speaker	3	3	--	--
c. Antenna and remote control patching systems	3	3	--	--
4. Purpose of components of a typical pulse radar set: Timer, modulator, transmitter, antenna, receiver, and indicators	3	--	3	--
5. Purpose of components of a typical sonar set: Automatic keying device, driver, transducer, receiver, and indicator	3	--	--	3
6. Operational capabilities and limitations of electronic equipment*	2	2	2	2

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Qualifications for Advancement in Rating	APPLICABLE RATES			
	ET	ETN	ETR	ETS
202 MAINTENANCE AND/OR REPAIR				
1. Effects of electrical shock, method of resuscitation of a person unconscious from electrical shock, and treatment for electrical and acid burns	3	3	3	3
2. Electrical and electronic safety precautions (except those applicable exclusively to line construction) as set forth in Chapter 18, U. S. Navy Safety Precautions (OPNAV 34P1)	3	3	3	3
3. Types of information shown and meaning of electrical, electronic, and mechanical symbols used in schematic diagrams of electronic equipment*	3	3	3	3
4. RMA color coding systems for: Fixed capacitors and resistors; power, AF, and IF transformer connections	3	3	3	3
5. System of assigning "AN" letter-number combinations as designation for electronic equipment	3	3	3	3
6. Define the following electrical, magnetic, and electronic terms:				
a. Volt	3	3	3	3
b. Ohm	3	3	3	3
c. Ampere	3	3	3	3
d. Watt	3	3	3	3
e. Volt-ampere	3	3	3	3
f. Henry	3	3	3	3
g. Farad	3	3	3	3
h. Cycle	3	3	3	3
i. Ampere-turn	3	3	3	3
j. Coulomb	3	3	3	3
k. Circular mil	3	3	3	3
l. Conductor and insulator	3	3	3	3
m. Field intensity	3	3	3	3
n. Flux density	3	3	3	3
o. Gauss	3	3	3	3
p. Permeability	3	3	3	3
q. Hysteresis and eddy current	3	3	3	3
r. Reactance, impedance, capacitance, and inductance	3	3	3	3
s. Self-inductance and mutal inductance	3	3	3	3
t. Electromagnetic induction	3	3	3	3
u. Power factor	3	3	3	3
v. Frequency	3	3	3	3
w. Phase	3	3	3	3
x. RC time	3	3	3	3
y. Attenuation	3	3	3	3
z. Absorption	3	3	3	3
aa. Conductance	3	3	3	3
bb. Modulation and demodulation	3	3	3	3
cc. Detection	3	3	3	3
dd. Selectivity and sensitivity	3	3	3	3
ee. Amplifier—class A, B, C, and AB	3	3	3	3
ff. Filter	3	3	3	3
gg. Intermediate frequency	3	3	3	3

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Qualifications for Advancement in Rating	APPLICABLE RATES			
	ET	ETN	ETR	ETS
202 MAINTENANCE AND/OR REPAIR—Continued				
hh. Heterodyne	3	3	3	3
ii. Node	3	3	3	3
jj. Resonance	3	3	3	3
kk. Nonlinear	3	3	3	3
ll. Sideband	3	3	3	3
mm. Zero beat	3	3	3	3
nn. AGC	3	3	3	3
oo. Ganged tuning	3	3	3	3
pp. Single side bands	3	3	3	3
7. Importance of using proper lubricants and solvents in maintenance of electronic equipment	3	3	3	3
8. Types and purposes of handtools and small portable power tools necessary for maintenance and repair of electronic equipment*	3	3	3	3
9. Function of the following test equipment:				
a. Nonelectronic multimeter	3	3	3	3
b. Electronic multimeter	3	3	3	3
c. Tube tester	3	3	3	3
d. Oscilloscope	3	3	3	3
e. Capacitance-inductance-resistance bridge	3	3	3	3
f. AF signal generator	3	3	3	3
g. RF signal generator	3	3	3	3
h. Frequency standards	3	3	3	3
i. Megohmmeter	3	3	3	3
j. Range mark recorder	3	--	3	3
k. Echo boxes	3	--	3	--
10. Methods and equipment used in electrical tests for continuity, grounds, and short circuits	3	3	3	3
11. Method of soldering and soldering equipment used in maintenance and repair of electronic equipment	3	3	3	3
12. Function of the following in electrical/electronic circuits: Resistors, rheostats, potentiometers, solenoids, inductors, relays, capacitors, fuses, switches, reactors, transformers, and crystals	3	3	3	3
13. Function of elements used in vacuum tubes	3	3	3	3
14. Construction of motors, generators, and alternators; application of laws of magnetism to electrical rotating machinery	3	3	3	3
15. Types, structure, maintenance procedures, and electrical characteristics of batteries	3	3	3	3
16. Relationship of resistance, inductance, and capacitance in a. c. circuits	3	3	3	3
17. Relationship of current, voltage, and impedance in a. c. circuits	3	3	3	3
18. Relationship of length and cross-sectional area to resistance of a conductor	3	3	3	3
19. Relationship of resistance, temperature, and current in an electrical conductor	3	3	3	3
20. Calculate current, voltage, and resistance in d. c. series and parallel circuits containing not more than four elements	3	3	3	3

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Qualifications for Advancement in Rating	APPLICABLE RATES			
	ET	ETN	ETR	ETS
202 MAINTENANCE AND/OR REPAIR—Continued				
21. Calculate current, voltage, phase angle, impedance, and resonance in a. c. series and parallel circuits containing not more than four elements	3	3	3	3
22. Function and operating principles of the following:				
a. Audio amplifiers	3	3	3	3
b. Video amplifiers	3	3	3	3
c. RF amplifiers	3	3	3	3
d. IF amplifiers	3	3	3	3
e. Oscillators—tickler -feedback, Colpitts, Ultra-audion, TPTG, push-pull, electron-coupled, transistron, Hartley-crystal controlled, and basic multivibrator	3	3	3	3
f. Rectifiers—copper oxide, selenium, crystal, and electron tube	3	3	3	3
g. Detectors—diode and crystal	3	3	3	3
h. Tuned coupling circuits	3	3	3	3
i. Impedance matching	3	3	3	3
j. Phase shifters	3	3	3	3
k. Cathode followers	3	3	3	3
l. Modulators—grid, screen, and plate	3	3	3	3
m. Limiters and clippers	3	3	3	3
n. AGC circuits	3	3	3	3
o. Sawtooth generators	3	3	3	3
p. Phase inverters	3	3	3	3
q. Modulation—amplitude, frequency, phase, and pulse	3	3	3	3
r. Coaxial transmission lines	3	3	3	3
23. Methods of obtaining three general types of bias: Fixed, cathode, and grid leak	3	3	3	3
24. Electrical characteristics of:				
a. Hertz, Marconi, and dipole antennas	3	3	3	--
b. Transducers and hydrophones	3	--	--	3
25. Cause and/or effect of sky wave, ground wave, ground-reflected wave, ionospheric reflecting layers, and skip distances	3	3	3	--
26. Methods of cleaning commutators and slip ring assemblies and precautions to be followed	3	3	3	3
27. Procedures for taking sonar beam patterns and range and bearing accuracy tests to indicate performance of transducers and equipment	2	--	--	3
28. Function of the following:				
a. Electronic switch	2	2	2	2
b. Synchroscope	2	--	2	2
c. Absorption wave meter	2	2	2	--
d. Grid dip meter	2	2	2	--
e. Radio-interference field-intensity meter	2	2	2	--
f. Spectrum analyzer	2	2	2	--
29. Operating principles of basic meters and meter movements employing the following:				
a. D'Arsonval and electrodynomometer meter movements	2	2	2	2

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Qualifications for Advancement in Rating	APPLICABLE RATES			
	ET	ETN	ETR	ETS
202 MAINTENANCE AND/OR REPAIR—Continued				
b. Shunts and multipliers	2	2	2	2
c. Thermocouples and rectifiers in a. c. meters	2	2	2	2
30. Method of connecting moving element to form voltmeter or ammeter; meaning of meter sensitivity and effect of circuit loading	2	2	2	2
31. Computations required to determine size of shunts and multipliers	2	2	2	2
32. Methods of performing sensitivity and selectivity measurements and alining circuits for optimum performance of electronic equipment*	2	2	2	2
33. Application of oscilloscopic wave form analysis to location of circuit malfunctions	2	2	2	2
34. Function and operating principles of vacuum tubes, gas-filled tubes, cathode ray tubes, and magnetrons	2	2	2	2
35. Operating principles of:				
a. Differentiators and integrators	2	2	2	2
b. Oscillators--blocking and Wein-bridge	2	2	2	2
c. Trigger circuits	2	2	2	2
d. Coincidence circuits	2	2	2	2
e. Peakers	2	2	2	2
f. Discriminators	2	2	2	2
g. AFC circuits	2	2	2	2
h. Counting circuits	2	2	2	2
i. Clampers	2	2	2	2
j. Phase splitting circuits	2	2	2	2
k. Paraphase amplifiers	2	2	2	2
l. Magnetic amplifiers	2	2	2	2
m. Transistors	2	2	2	2
36. Applications and operating principles of:				
a. Wave guides	2	2	2	--
b. T/R and AT/R tubes	2	--	2	--
c. Klystrons and magnetrons	2	--	2	--
d. Crystal mixers	2	2	--	--
e. Radar modulators	2	--	2	--
37. Cause and/or effect of induction field and radiation field	2	2	2	--
38. Characteristics and use of synchros and servomotors; methods of setting to electrical zero; purpose of gain, phase, and balance adjustments	2	2	2	2
39. Operating principles and radiation characteristics of parabolic and lens antennas	1	1	1	--
40. Operating principles and characteristics of, and repairs and calibrations authorized on electronic test equipment listed in items 202.9 and 202.28	1	1	1	1
41. Function and operating principles of special purpose tubes such as: Traveling wave, carcinotrons, and high-powered klystron amplifiers	1	1	1	1

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Qualifications for Advancement in Rating	APPLICABLE RATES			
	ET	ETN	ETR	ETS
202 MAINTENANCE AND/OR REPAIR—Continued				
42. Effects of environmental conditions upon operation of electronic and electrical equipment and special maintenance techniques involved for equipment to be operated at extremes of temperature and humidity	C	C	C	C
43. Theory of single side band radio transmission and reception	C	C	C	C
44. Polarization and directional characteristics of antenna arrays such as: Driven arrays (collinear), parasitic arrays (Yagi), parabolic, corner or flat reflectors, phased arrays, wage guide, and type antennas	C	C	C	C
203 ADMINISTRATIVE AND/OR CLERICAL				
1. Types of information contained in electronic technical and maintenance publications	3	3	3	3
2. Types of entries and information recorded in electronic equipment failure reports, work logs, equipment histories, checkoff lists, and Current Ship's Maintenance Project (CSMP)	3	3	3	3
3. Procedures for taking, recording, and reporting inventories of tools and portable test equipment used in maintenance and repair of electronic equipment	3	3	3	3
4. Procedures for obtaining replacement parts and supplies	2	2	2	2
5. Types of information included in job orders and work requests	2	2	2	2
6. Types of information reported in periodic or recurring reports concerning performance and/or maintenance of electronic equipment	1	1	1	1
7. Methods, techniques, and devices applicable in electronic maintenance training of teams and individuals	1	1	1	1
8. Procedures for organizing and administering the maintenance and repair of electronic equipment	C	C	C	C
9. Current bureau directives on operational and maintenance responsibilities of enlisted personnel for electronic equipment	C	C	C	C
10. Procedures for accounting for electronic equipment, maintaining control of inventories and work flow, and reporting equipment status and work accomplished	C	C	C	C

300 PATH OF ADVANCEMENT TO WARRANT OFFICER AND LIMITED DUTY OFFICER

Electronics Technicians advance to Warrant Electronics Technicians and/or to Limited Duty Officer, Electronics.

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